ATS 64096

N73 18178

ELECTRONICS LABORATORY • SYRACUSE, N.Y.

Final Report

LOW COST GROUND RECEIVING SYSTEMS FOR TELEVISION SIGNALS FROM HIGH-POWERED COMMUNICATIONS SATELLITES

VOLUME I

NASA———CR 120933

J. P. Hesler
Y. C. Hwang
J. J. Zampini

GENERAL ELECTRIC

Final Report

LOW COST GROUND RECEIVING SYSTEMS FOR TELEVISION SIGNALS FROM HIGH-POWERED COMMUNICATIONS SATELLITES

VOLUME I

NASA——CR 120933

J. P. Hesler
Y. C. Hwang
J. J. Zampini

Prepared by

Electronics Laboratory
GENERAL ELECTRIC COMPANY
Syracuse, N. Y.

TABLE OF CONTENTS

TABLE OF CONTENTS (concluded)

LIST OF ILLUSTRATIONS

LIST OF ILLUSTRATIONS (continued)

LIST OF ILLUSTRATIONS (concluded)

LIST OF TABLES

LIST OF TABLES (continued)

1.0 INTRODUCTION

The purpose of the program described herein, performed under NASA Contract NAS-3-11520, was to investigate and fabricate ground signal processing systems for television satellites. This work is a part of the general investigation of satellite television systems conducted by NASA.

The program consisted of nine original tasks, divided into two phases, plus several additional tasks that concerned special subsidiary subjects, as outlined below. The results of the first four tasks, compromising Phase I, were reported in a summary report.(1) This final report covers the remainder of the work performed under the contract.

<u>Phase I</u>

TASK I	Candidate System Selection
TASK II	Subsystem Performance Evaluation
TASK III	Cost Screening
TASK IV	Conceptual Design and Optimization

<u>Phase II</u>

TASK V	Design, Breadboarding, Test and Evaluation (partially completed in Phase I)
TASK VI	Production Design, Testing and Performance Evaluation
TASK VII	Fabrication and Acceptance Testing
TASK VIII	Shock and Vibration Testing
TASK IX	Program Management (All Phases)

<u>Target Cost Converter Study</u>

TASK X	Target Cost Converter, Candidate System Selection, Analysis and Costing
TASK XI	Breadboard Development and Evaluation
TASK XII	Prototype Fabrication and Evaluation

---

(1) Ground Signal Processing Systems, Summary Report on Analysis, Design and Cost Estimating. NASA CR-72709; June 1970.

Cost Sensitivity Analysis

TASK XIII Analysis of the Converter Factory Costs as Functions of Input Signal Level and Output S/N Ratio

Integrated Tunnel Diode Amplifier Development

TASK XIV Design and Fabrication of a 12 GHz Tunnel Diode Preamplifier.

Ten, each, converters, of three types, were fabricated and delivered in Phase II. The converter types differed with respect to transmission carrier frequency and modulation format. All were designed to process a standard NTSC color television signal and provide outputs compatible with unmodified NTSC color television receivers. A summary of the major format parameters is listed in Table I.

The Target Cost Converter contract modification was aimed at determining if a substantial production cost saving could be realized by designing a 12 GHz FM converter within a given factory cost constraint while relaxing design requirements and objectives.

The Cost Sensitivity Analysis task was performed to determine converter factory costs as functions of available input signal power and various output signal to noise ratios. This study was performed for each converter type.

The integrated tunnel diode amplifier task was directed toward the development and fabrication of a low cost 12 GHz tunnel diode amplifier that would provide a total satellite television system cost advantage.

TABLE I

FORMAT OF CONVERTER TYPES

CONVERTER TYPE	A	B	D
Carrier Frequency (GHz)	2.25	2.25	12.00
Video Modulation	AM/VSB	FM	FM
FM Modulation Index		2	3
Aural Modulation	4.5 MHz FM Subcarrier referred to video baseband		
Nominal Input Signal Power (dBw)	-92	-107	-105
Nominal RF Bandwidth (MHz)	6	30	40
Intermediate Frequency (MHz)	85	120	120
Output Carrier Channel	6	5/6	5/6

2.0 SUMMARY OF RESULTS

The results of the Phase I study indicate that low cost converters could be designed and produced to meet the needs of a large consumer market in a synchronous satellite television system. Candidate converter configurations were selected for each type in a cost optimization procedure. Preliminary breadboard verification of key circuit elements,as well as detailed cost analysis of the candidate designs, was performed. Based on these Phase I results the NASA Contract Manager authorized the Phase II effort, which resulted in the fabrication and evaluation of 10 engineering prototype units of each of the three converter types.

Each of the converter designs consisted of two units, an antenna-mounted unit and a receiver-mounted unit. The ground receiving antenna was not specified as a portion of the design problem in this contract. Photographs of the converter units are shown on pages 4, 5, and 6.

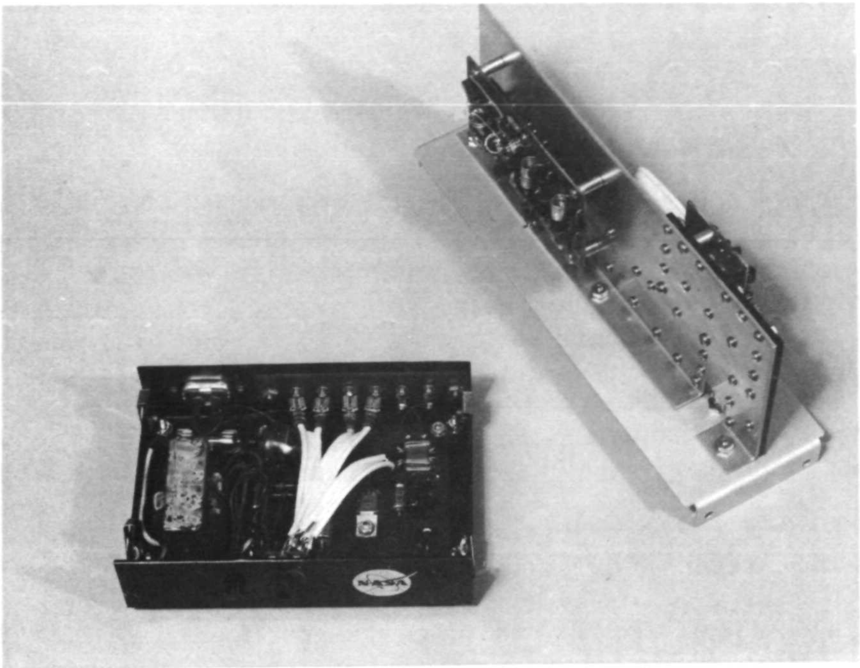
The sizes and weights of the units are listed in Table II.

TABLE II.

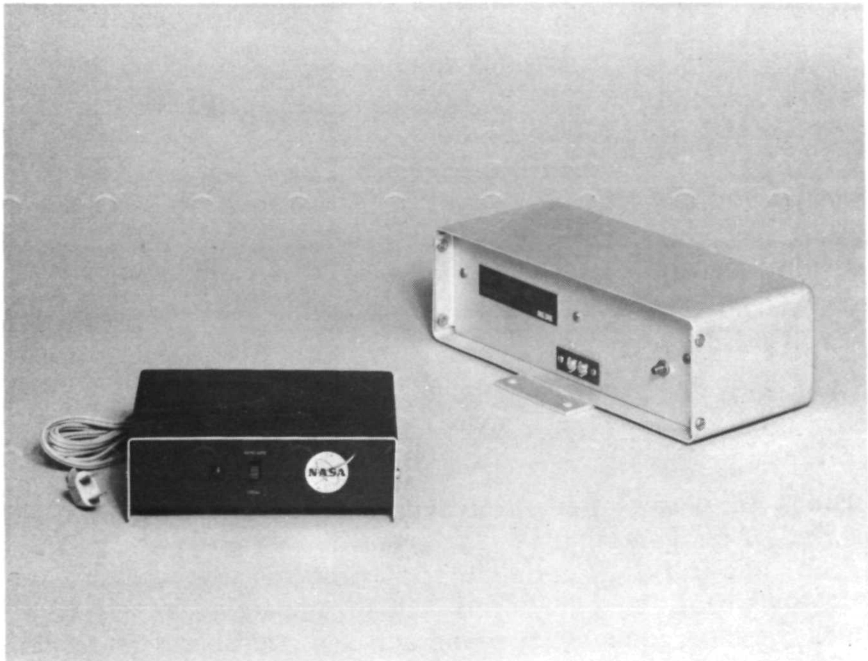
PHYSICAL CHARACTERISTICS OF CONVERTER TYPES

	Dimensions							
Type A S-Band AM	L		W		H		Weight	
	cm.	in.	cm.	in.	cm.	in.	kg.	oz.
Antenna Unit	26.7	10.5	8.9	3.5	16.3	6.4	0.91	32
Indoor Unit	16.5	6.5	11.9	4.7	5.5	2.15	0.76	27
Type B S-Band FM								
Antenna Unit	13.3	5.25	9.1	3.6	14.6	5.75	0.40	14
Indoor Unit	24.8	9.75	12.1	4.75	6.4	2.5	1.14	40
Type D X-Band FM								
Antenna Unit	12.7	5.0	8.1	3.2	13.3	5.25	0.62	22
Indoor Unit	24.8	9.75	12.1	4.15	6.4	2.5	1.30	46

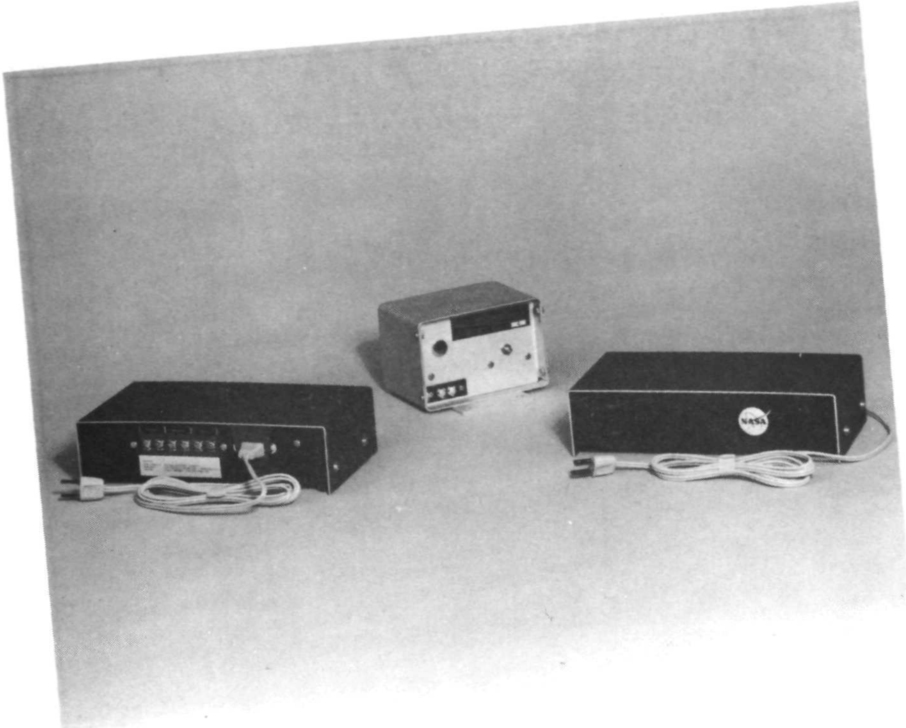
Block diagrams of the prototype units are shown in Figures 1, 2 and 3. Each type of converter uses a balanced mixer front end. The mixer, local oscillator and initial IF amplification circuits are contained in the antenna units. Providing the microwave to IF frequency translation in the antenna units permits the interconnection of the antenna and receiver mounted units, specified as 30 feet of cable, at a frequency where the cable loss is tolerable.



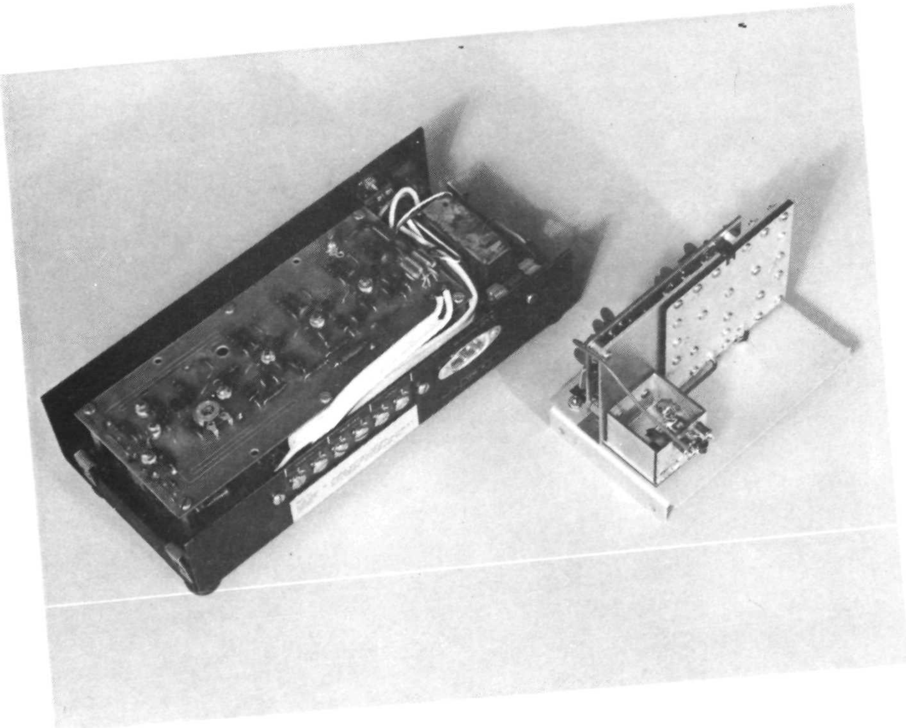
Type A Converter
2.25 GHz AM



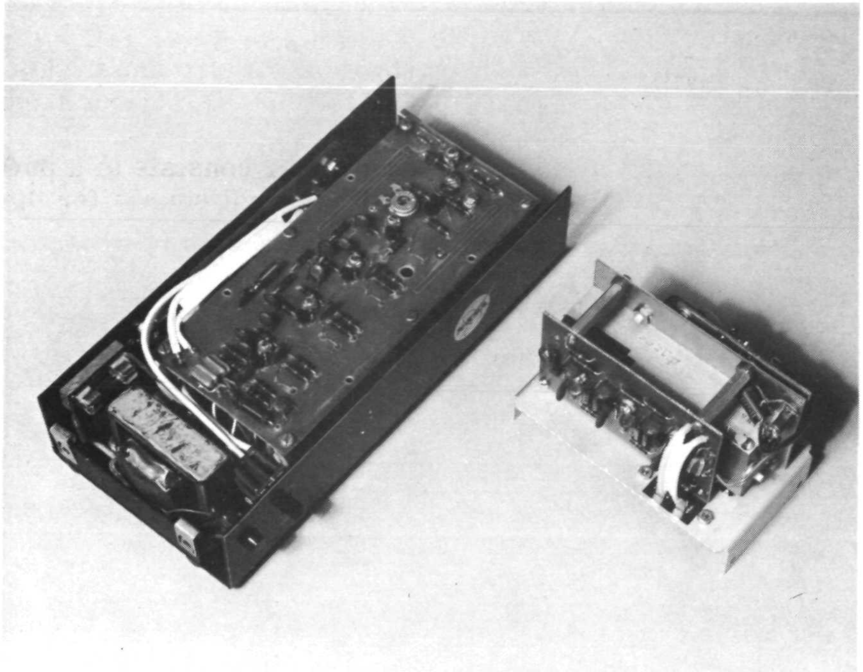
Type A Converter
2.25 GHz AM



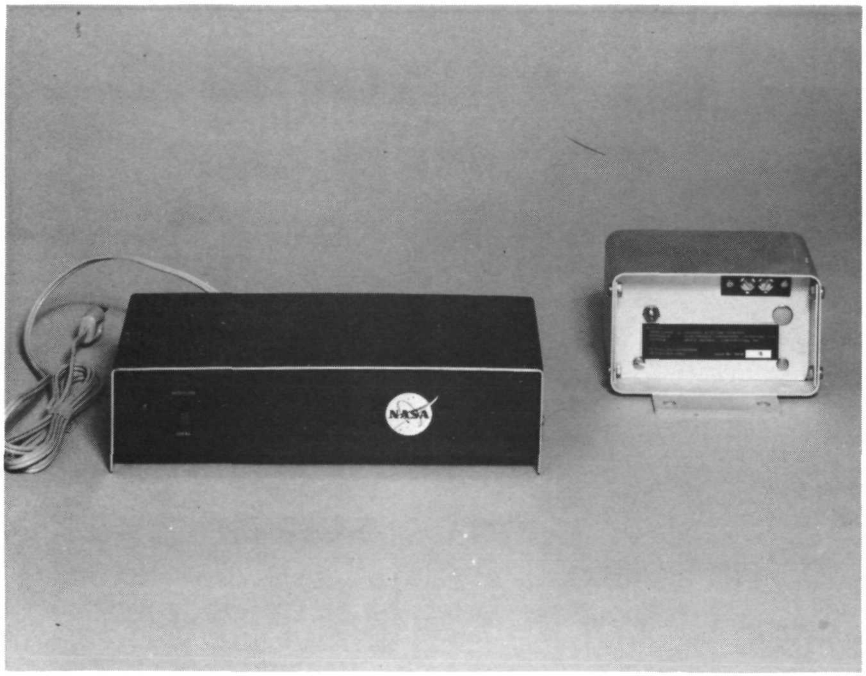
Type B Converter
2.25 GHz FM



Type B Converter
2.25 GHz FM



Type D Converter
12 GHz FM



Type D Converter
12 GHz FM

The Type A, S-band AM Converter translates the microwave signal directly to channel six in the VHF television band. Only one stage of IF amplification is required to establish the converter noise figure and in this case the intermediate frequency coincides with the channel six carrier frequency.

The indoor unit of the Type A converter consists of a power supply and an antenna/power switch that selects the signal source for the television receiver from either the microwave converter or the local VHF antenna. This switch also removes the line power from the converter power supply when the microwave source is not selected.

The Type B converter receives an S-band FM signal and translates it to a 120 MHz IF in the antenna unit. The indoor unit for this converter provides additional IF gain and demodulates the FM signal to recover a baseband video signal and an FM audio subcarrier at 4.5 MHz. The composite video and audio signal is remodulated, AM/DSB, at the VHF channel five or six frequency to interface with the standard unmodified color television receiver.

The indoor unit for the Type B converter also contains the dc power supplies and antenna/power switch, which serve the same function as in the Type A converter.

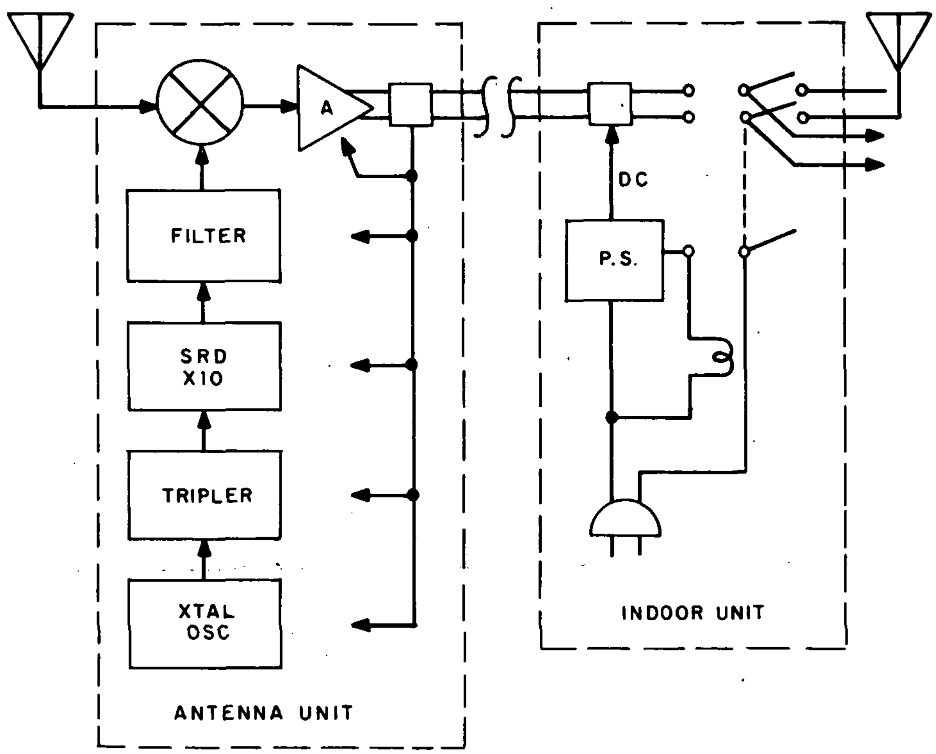


Figure 1. Type A Converter, Block Diagram

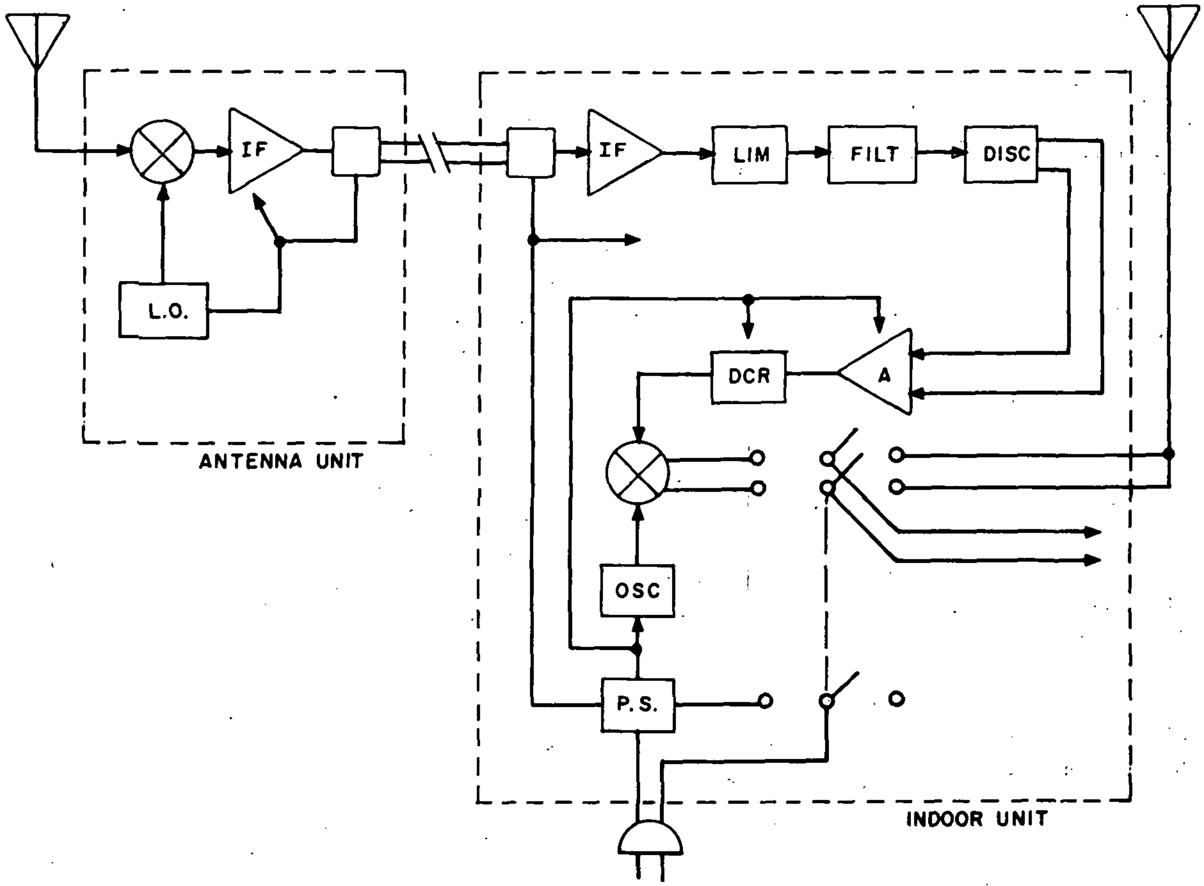


Figure 2. Type B Converter Block Diagram

The Type D converter is also an FM receiver and its functions are identical to the Type B converter. The mixer and local oscillator circuits are different, to accommodate the X-Band FM carrier, and the power supply regulator details are changed to fit the circuit needs.

Some circuit changes have been made subsequent to the cost analyses performed in Phase I and published in the summary report. Modified factory cost estimates are provided in Table III to account for these circuit changes and new price data on component parts.

Noise figure measurements for the FM converters are listed in Table IV. These values are wideband noise figures measured at the output of the antenna unit IF amplifier. FM noise from the local oscillators has not been accounted for in these measurements. Independent evaluation of the FM noise characteristics of the local oscillators has shown that this source of noise is negligible in the FM converters.

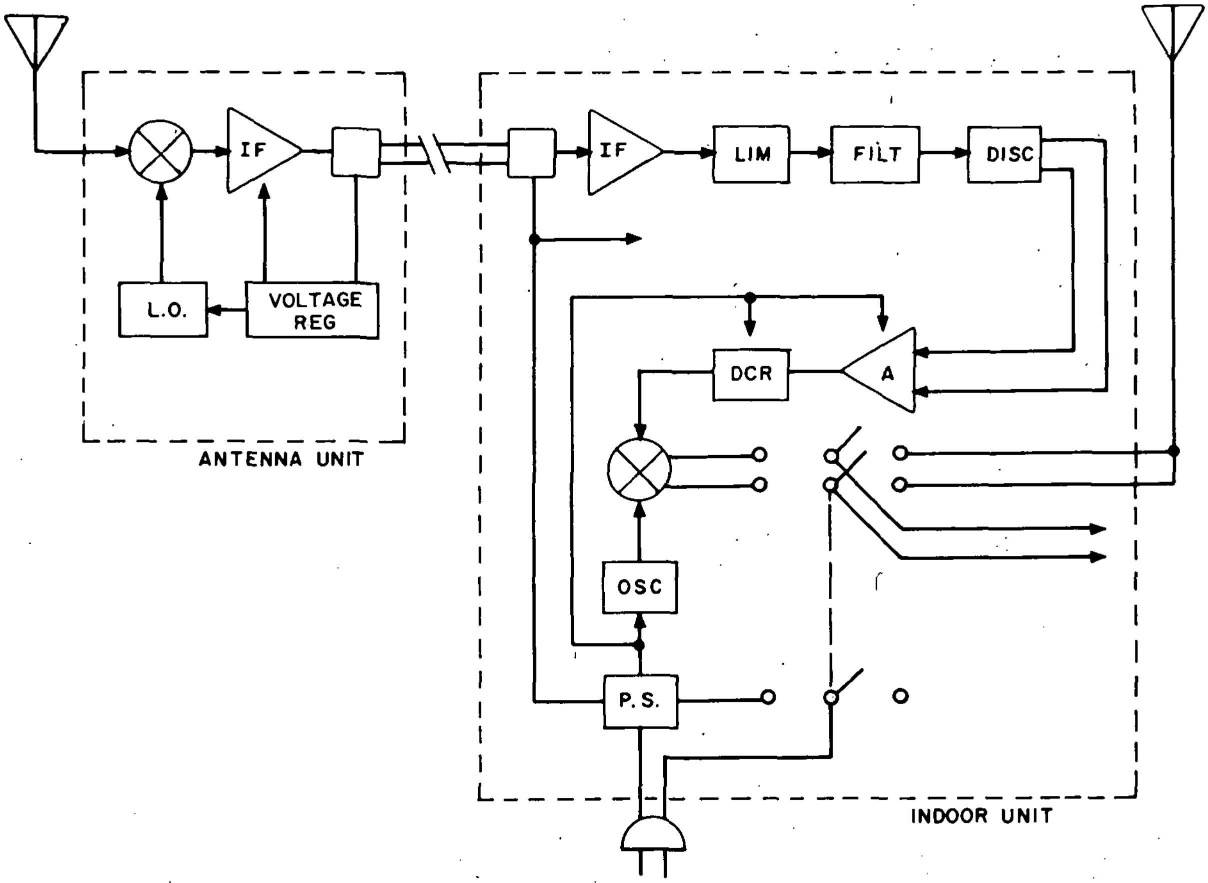


Figure 3. Type D Converter Block Diagram

The double side band noise figures were measured by adding excess noise to the receiver inputs until the IF output power was doubled. The operating noise figures for the Type B converters will be one to two dB greater than indicated, instead of 3 dB greater, because of the attenuation of image noise by the image filters in the 2.25 GHz mixers. The Type D converter operating noise figures will be 3 dB higher than indicated because no image filters are used in the 12 GHz mixers.

The noise figure of the Type A converters was not measured directly because the local oscillator multiplier chain produces a spurious output within the IF amplifier passband at the local oscillator crystal frequency.

All converter antenna units were operated over the specified temperature range of -40°C to 55°C. Representative samples of the indoor units were operated over the small temperature range, -1.1°C to 60°C, specified for the indoor units. A major portion of the development effort was concerned with the frequency stabilization of local oscillator circuits over the

TABLE III

UPDATED CONVERTER FACTORY COST ESTIMATES (1970 COSTS)

TYPE A CONVERTER, 2.25 GHz AM

10^3	10^4	10^5	10^6
32.31	23.56	20.40	18.35

TYPE B CONVERTER, 2.25 GHz FM

42.34	28.80	24.68	21.83

TYPE D CONVERTER, 12.0 GHz FM

83.92	59.75	40.37	32.78

TABLE IV

FM CONVERTER NOISE FIGURE MEASUREMENTS

Type B 2.25 GHz		Type D 12 GHz	
SFM-1	7.8 dB	XFM-1	*
SFM-2	8.9	XFM-2	*
SFM-3	7.3	XFM-3	8.0
SFM-4	10.6	XFM-4	7.8
SFM-5	7.8	XFM-5	8.7
SFM-6	8.2	XFM-6	8.4
SFM-7	8.1	XFM-7	8.0
SFM-8	7.0	XFM-8	9.9
SFM-9	6.7	XFM-9	7.8
SFM-10	8.0	XFM-10	7.3

* Inoperative Due to Gunn Diode Shortage

specified antenna unit temperature range. The greatest problems occurred at the lower temperature limits, i.e., below -20°C, with the crystal oscillator and frequency multiplier chain used in the Type A system.

All design requirements for the converters were met with the prototype units.

The design objectives were met with selected units of each type as discussed below.

No intermodulation products were visible in television receiver output using the Type A system. The AM converter was designed as a relatively wideband converter, and channel bandwidth or selectivity is provided by the television receiver. The normal audio subcarrier - color subcarrier intermodulation signal can be observed when the television receiver is mistuned; however, these intermodulation products are produced within the television receiver.

The same intermodulation distortion, sound-color, can be produced using the FM converters when the IF and discriminator circuits in the converters are misaligned. With proper alignment, this source of intermodulation distortion is barely discernible in the television picture. The performance goal — a negative 40 dB intermodulation product amplitude — was exceeded in the FM converter systems.

Each converter type was designed to minimize the influence of the converter on video signal bandwidth and delay distortion. The AM converter is wideband with respect to a six megacycle channel and no distortion effects are evident.

The video bandwidth of the FM converter demodulators is also wider than required. The FM demodulator/video amplifier is slightly peaked around 4.5 MHz. The differential delay out to 4.5 MHz is less than 20 nanoseconds referred to the low frequency envelope delay. The remodulator bandwidth is also in excess of that required. The result is that the television receiver establishes the video noise bandwidth in the system.

The design objectives for the output signal to noise ratios for the various converter types were as follows:

Type A, 35 dB S/N for -92 dBW antenna output power level

Type B, 35 dB S/N for -107 dBW antenna output power level

Type D, 35 dB S/N for -105 dBW antenna output power level

The design objectives were met with selected units of each type. The Type D converters provided the best performance followed by the Type B and Type A converters in that order.

A switch is used in each converter type to select between the converter output and the local VHF television antenna as a signal source. An inexpensive DPDT slide switch, which has low line-to-line capacitance, is used. The signal power attenuation, using this switch in a balanced 300Ω twinlead system, is less than one decibel. The converter power is switched off when the local VHF antenna is selected for television reception to minimize interference.

2.1 TARGET-COST CONVERTER

The target-cost converter design effort was initiated to determine the possible cost-performance tradeoffs that could be realized by specifying cost objectives rather than performance objectives. The type D, X-Band FM, converter was selected as a vehicle. A target cost was chosen that was 50 percent of the factory cost estimate arrived at in Phase I for the Type D converter. Design requirements were relaxed to permit integration of the converter and television receiver, and performance objectives were relaxed.

The first task of this program, "Candidate System Selection, Analysis and Costing", was completed. Concurrently a portion of the second task, "Breadboard Development and Evaluation" was performed.

Several methods for reducing the converter cost under the new ground rules were considered and evaluated. These ideas were focused on the relatively high cost areas of X-Band mixers and local oscillators. A single diode mixer breadboard was fabricated and tested using microstripline construction. Tests were made to determine the feasibility of using a Gunn oscillator in a dual role of oscillator and mixer. Elimination of the indoor unit as a separate hardware item by building the indoor circuit functions into the television set was investigated.

The feasibility studies and cost analyses indicated; 1) no practical converter could be realized at the target cost selected; 2) the configuration of a converter that would provide reasonable performance would not be significantly different in concept from the Type D converter, and 3) the cost savings resulting from a minimal performance design did not justify the performance degradation.

The program was terminated at the midpoint and no target-cost converter prototypes were fabricated.

This study demonstrated that the "target cost" theory of product cost/performance optimization is not applicable in a product that has limited performance tradeoff capacity. In this case, there was a very small range of converter RF sensitivity available to work with. In addition, the threshold characteristics of FM receivers further limit the practical range of IF carrier to noise operation.

2.2 COST SENSITIVITY ANALYSIS

The cost sensitivity analysis task provided cost estimates for each converter type as a function of two operational parameters. The first parameter was RF input signal level; the second, output signal to noise ratio. Converter configurations were selected to minimize cost for each set of parameters.

The results of this test are discussed later in Section 5.0.

3.0 CONVERTER CONFIGURATIONS

This section of the report contains technical discussions pertaining to design options and circuit details presented in the Phase I Summary Report. Detailed technical substantiation and documentation are provided in Volume II. The converters will be considered in sequence; Type A, Type B and Type D.

3.1 TYPE A CONVERTER

3.1.1 General

The Type A converter is distinctive for two basic reasons. First, the amplitude modulation transmission format is directly compatible with the NTSC color receiver except for the carrier frequency. No signal detection is required in the converter, only frequency translation. Second, the television channel frequency allocations have minimum guard band and, therefore, the television receivers have a minimum requirement for fine tuning and automatic frequency control. This is reflected in the Type A converter design by a low drift requirement for the microwave local oscillator. The allowable L.O. drift is limited by the television receiver fine tuning and AFC capabilities. For this reason a stable local oscillator design is required, and stability is required over the temperature range of $-40^{\circ}C$ to $+55^{\circ}C$ specified for the antenna unit. A design goal of ± 250 kHz L.O. drift over the temperature range was selected. This stability requires either stable crystal reference or an AFC circuit in the converter. Since the converter did not require the amplification to reach a detectable signal level, an AFC function was not available in the basic converter circuitry. AFC from the television receiver was also unavailable because access to the television receiver circuits was prohibited by the design constraints. For these reasons a crystal oscillator and multiplier chain was selected for the Type A converter. The crystal oscillator frequency was limited by the cost of crystals as a function of frequency. The crystal frequency was selected as one-thirtieth of the desired microwave local oscillator frequency or 72.225 MHz.

$$f_{crystal} = \frac{f_T - f_{CH6}}{30}$$

The oscillator circuit finally selected for this converter is shown in Figure 4. The previous oscillator circuit used the crystal as a series feedback component between the oscillator transistor collector and base. This introduced a dc bias on the crystal and caused starting problems at low temperature. The present circuit eliminates the crystal bias without adding component parts and improves the overall circuit operation. A resistor, R_5, was added between the oscillator and the oscillator buffer amplifier to reduce loading effects on the crystal oscillator.

A substantial design and evaluation effort was applied to the crystal oscillator and tripler circuits to provide a temperature stable design. A fourth transistor was added to the chain to permit the use of stabilizing circuits while retaining the same overall power gain. The final tripler and tripler buffer design uses class C amplifiers as shown in Figure 5.

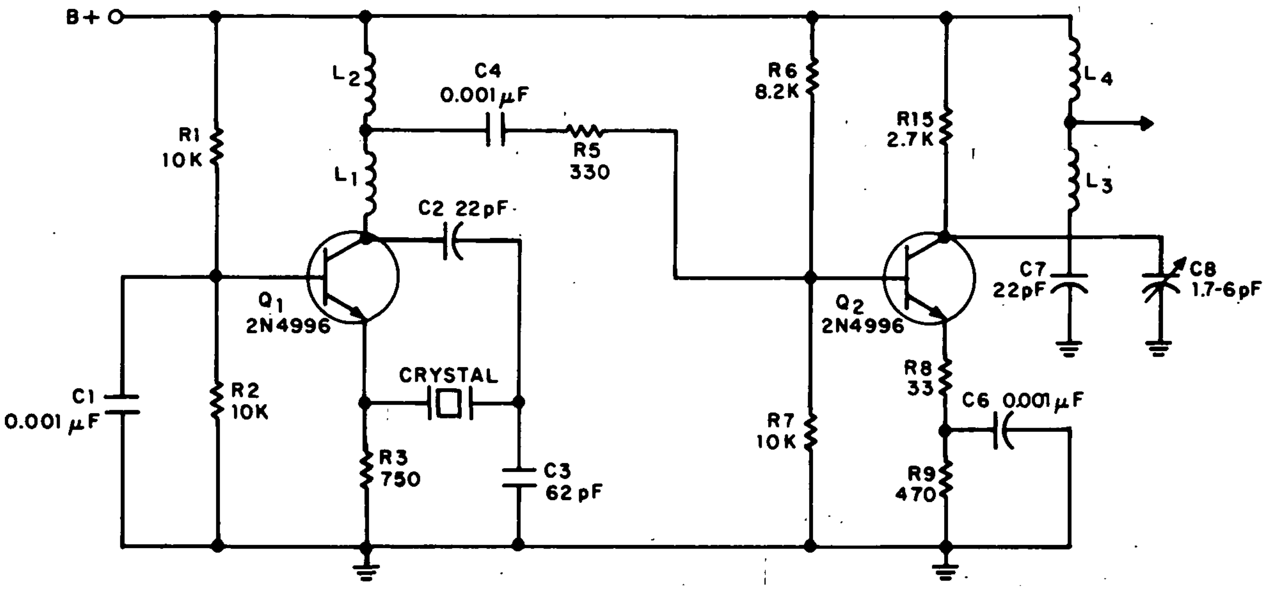


Figure 4. Type A Converter, Local Oscillator, Circuit Crystal Oscillator and Buffer Stages

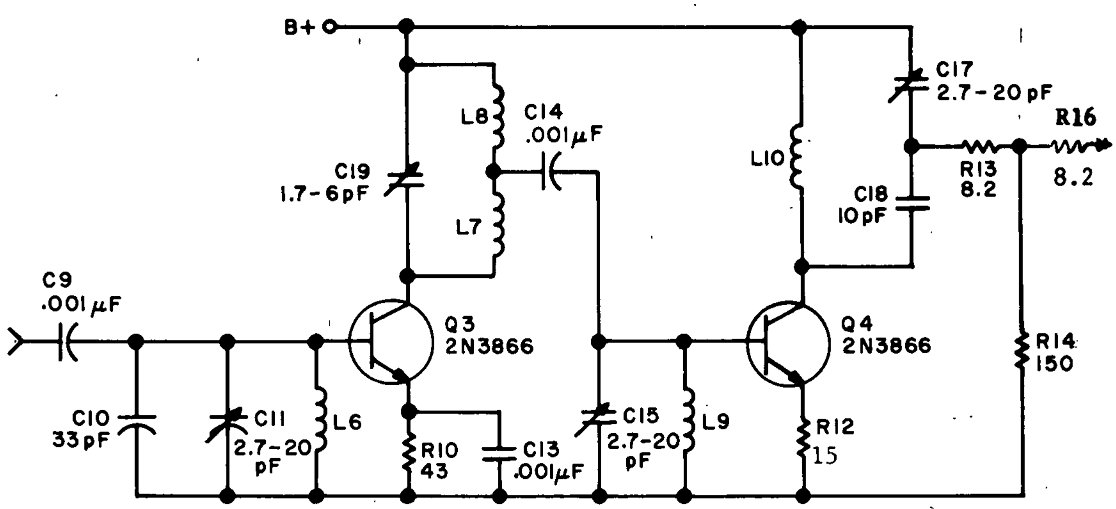


Figure 5. Type A Converter, Local Oscillator Circuit, Tripler and Buffer Stages

Parallel tuned tanks are used in the base and collector circuits of these stages. The tripler has a bypassed emitter resistor for conduction angle control. Tuning is provided with five variable capacitors on the oscillator tripler board. The interstage networks have been reduced to a minimum to facilitate alignment and to reduce cost.

The tripler buffer circuit drives the times-ten step recovery diode multiplier circuit at a power level of approximately 100 milliwatts. The interface of these two circuits presented a problem. Each circuit passed the temperature tests when tested independently with dummy sources and/or dummy loads. When the SRD multiplier was driven with the tripler it was found that reflected impedance variations with temperature changes resulted in malfunctions. A resistive T-pad having six dB return loss was inserted between the circuits with an appropriate increase in the tripler buffer output level and this decoupling solved the problem. Some further low temperature malfunctions in the prototype tripler units was traced to low gain 2N3866 transistors. It is, therefore, necessary to screen these units for a minimum h_{FE} of 40 prior to assembly.

3.1.2 X10 SRD Multiplier

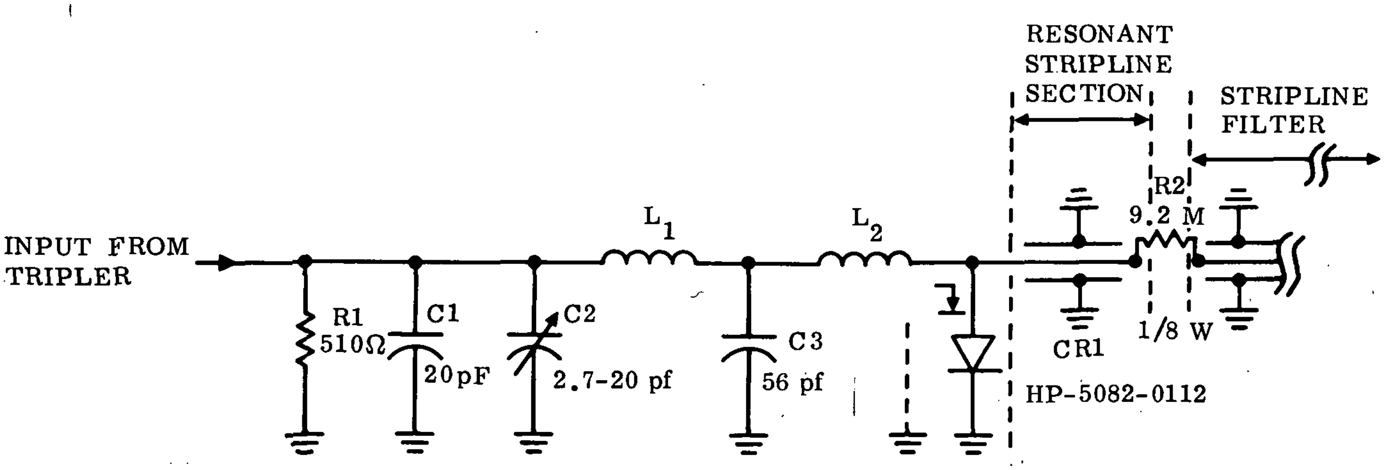
The SRD multiplier consists of a step recovery diode, an input matching circuit, and a resonant stripline section.

The step recovery diode multiplier circuit used in the prototype converters is almost identical to the circuit shown in the Phase I report. The variable resistor in the self bias network was replaced with a fixed resistor. One tuning capacitor is provided for alignment. The SRD multiplier drives a resonant section of stripline, which is lightly coupled to a stripline filter. The coupling is provided through the capacitance of a 1/8 watt carbon resistor that is connected in series between the stripline resonant section and the stripline filter.

The circuit diagram for the SRD multiplier is shown in Figure 6. The input to the times-ten multiplier comes from the T-pad at the output of the oscillator tripler circuit. This pad provides a 6 dB return loss and isolates the tripler and SRD multiplier circuits.

3.1.3 SRD Multiplier Filter

The stripline filter selects the desired harmonic from the spectral output of the SRD multiplier. In this case the desired frequency is 2166.75 MHz ±250 kHz. It is not practical to design a stripline filter with extremely narrow bandwidth because of etching tolerances and temperature effects. The stripline filter was designed to select the tenth harmonic of the 216.67 MHz SRD drive frequency. The bandwidth of the filter was selected to be a compromise between unwanted harmonic suppression and producibility at low cost without tuning adjustments. The filter that follows the multiplier chain is a five section filter using half-wave line sections that do not require shorting at the ends. The resonant line section is excited by the sharp current transients generated in the step recovery diode at a repetition rate equal to the drive frequency. The transient, being rich in harmonic content, contains



L_1 = 1 3/4 TURNS, 0.018" DIAMETER WIRE, 0.18" I.D.

L_2 = 0.2 ", 0.018" DIAMETER WIRE

Figure 6. SRD Multiplier Circuit

ample energy at 10 times the drive frequency, to act as the desired local oscillator signal. The resonant line section is lightly coupled to the output stripline filter to preserve its Q. The desired harmonic of the SRD drive signal is enhanced by the bandpass characteristic of the resonant line section. The circuit is broadband, however, and the energy at the adjacent harmonic frequencies is also large. The desired harmonic is selected by the stripline bandpass filter while the adjacent harmonics are suppressed.

Although this filter is larger than an equivalent filter using quarter wave shorted sections, it is easier to fabricate and it has more predictable performance. The bandpass response of this filter, as shown in Figure 7, extends from 2134 MHz to 2180 MHz at the -3 dB power points relative to midband response. This filter suppresses the adjacent ninth and eleventh harmonics of the SRD multiplier by at least 50 dB. The filter passes some of the sideband energy at $f_{LO} \pm f_{crystal}$. This energy is present as a result of imperfect filtering in the tripler chain which leaves some residual amplitude modulation on the SRD drive at the crystal oscillator frequency. The SRD multiplier, being a wideband circuit, responds to amplitude variations in its drive signal by producing amplitude variations in its output spectrum. Some of the amplitude modulation sideband energy at frequencies of

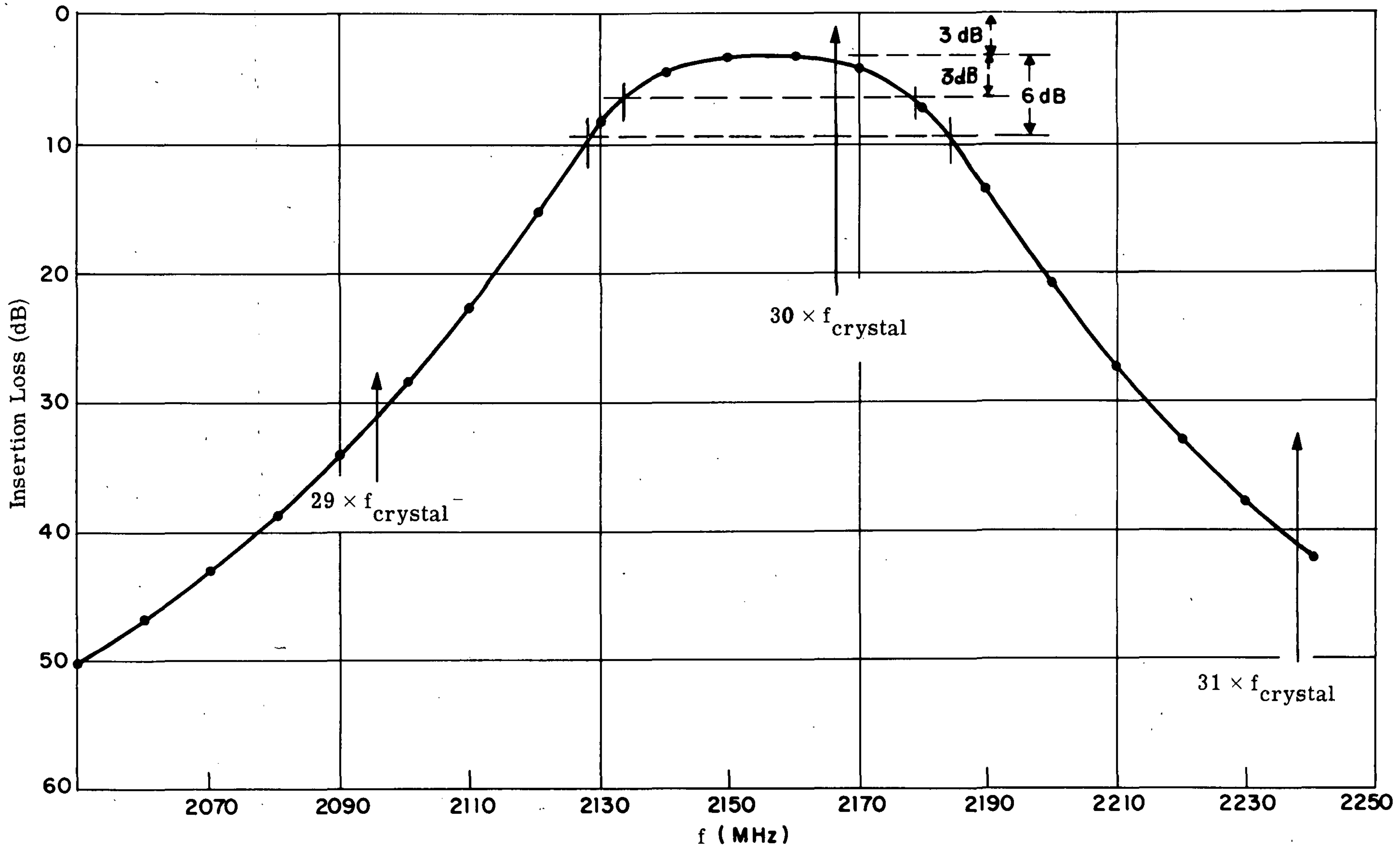


Figure 7. Type A Converter, SRD Bandpass Filter Response

$3 \times 10\ f_{crystal} \pm f_{crystal}$ appears at the mixer input along with the desired local oscillator power of $3 \times 10\ f_{crystal}$. The mixer treats this energy as any other signal and translates it to the difference frequency, $f_{crystal}$. The IF amplifier, therefore, sees the sideband energy leakage that passes through the post multiplier filter. The level of this leakage signal is equal to or greater than the desired signal power in the IF because of the wide disparity between the RF signal level and the local oscillator power level. This presents no problem in the Type A converter as long as the IF amplifier or television receiver front end is not overloaded with this undesirable signal. The same is not true for the Type B converter and this type of unwanted signal interfered with the FM signal demodulation. The situation presents no problem in the AM converter unless the tripler circuit is grossly misaligned.

3.1.4 Mixer

The S-Band mixer used in the Type A system is the same balanced mixer with hybrid feed discussed in the previous report.* The mixer layout was modified to include image rejection filters on both the RF and L.O. ports. The purpose of the image rejection filters is to increase the mixer conversion loss in the band $f_{L.O.} - f_{IF}$ and, thereby, improve the converter noise figure. The image filter reduces the converter noise figure by one to two dB.

A terminating resistor was added at the output of the SRD multiplier filter to improve the mixer performance. An 83 ohm, 1/8 watt resistor is placed in shunt with the SRD multiplier filter output within the stripline sandwich. The test results indicate that a mismatch condition still remains in the mixers of certain prototype converters and the problem has not been fully diagnosed.

The diodes used in the mixer for the Type A system are axial lead Schottky diodes, which are the least expensive Schottky diodes available. These diodes were selected because the cost increment was significant for diodes that provide improved performance. The diodes were selected as matched pairs using a curve tracer before being assigned to a particular mixer in production of the prototype units. Improved performance could be obtained at additional cost by using beam-leaded Schottky diodes in the S-band converters.

PPO[(1)] printed circuit board material was used for the S-Band mixers and the stripline filter in the Type A system multiplier chain. Two other materials were also evaluated, Rexolite (TM) and Florglass (TM). The Rexolite (TM) material was found to be unsatisfactory for two reasons. First, excessive warping of the material was encountered when a large percentage of the copper was removed from one side of the double-clad material required in the stripline fabrication. This is a result of stress in the dielectric material, introduced in the lamination process of the double-clad board. A similar problem was encountered with PPO material that was not annealed. The second problem with the Rexolite (TM) was a tendency to craze when the board was etched, and to crack where the warping was removed during circuit assembly.

---

*Drawing SK56157-D91-73. See page 148.

(1) PPO, polyphenylene oxide.

The Florglass (TM) became available during the program and was found to be as good as, or superior to, the PPO material. The Florglass (TM) is constructed using teflon dielectric with glass fiber filling to maintain structural rigidity and minimize cold flowing. Teflon has a higher melting point than the PPO and is, therefore, easier to use where soldered connections are required. The dielectric constants of the PPO and Florglass (TM) are near enough to permit direct substitution without artwork changes.

The mixer and SRD output filter are fabricated in a continuous stripline circuit to minimize connections. This causes the length of the Type A converter antenna unit package to be the largest of all converter types.

3.1.5 IF Amplifier

A circuit modification was made to reduce parts count in the IF amplifier for the Type A circuit as shown in Figure 8. A balun is used, as before, to couple the IF signal and bias power between the antenna unit circuits and the twinlead cable. In the modified circuit the transistor amplifier is connected as a common emitter stage and the balun is a one-to-one impedance ratio coupler to the balanced twinlead.

3.1.6 Indoor Unit, Type A Converter Power Supply

The Type A converter power supply is shown in Figure 9. The Type A converter uses a regulated power supply derived from a transformer coupled full-wave rectifier. A center-tapped transformer is used to save two rectifier diodes. The transformer primary is fused and an RFI filter is provided on the line cord entrance. A neon pilot light is connected across the transformer primary and the primary power is switched with a SPDT segment of the Antenna switch.

The voltage regulator uses a power Darlington transistor as the series regulating element. The error voltage is sensed by a Zener diode, which connects the regulated output voltage to the base of a feedback transistor, Q-2. When the output voltage reaches the Zener voltage plus, the V_{BE} drop of the feedback transistor, a portion of the Darlington transistor drive current is shunted through the feedback transistor. The high current gain of the Darlington transistor permits the feedback circuit to operate at a low current level. A filter capacitor from the base of the Darlington transistor to ground stabilizes the feedback loop. A series dropping resistor in the collector circuit of the Darlington protects the power supply from inadvertent shorts in the twinlead connection to the antenna unit.

The Zener diodes are selected entertainment grade silicon transistors having the desired emitter-base breakdown characteristics. The collector and base terminals are tied together in the Zener application.

3.1.7 Type A Converter Packaging

The Type A converter circuits are housed in two separate packages, the antenna unit, and the indoor unit. The antenna unit circuits, local oscillator, mixer, and IF amplifier, are housed in a deep drawn

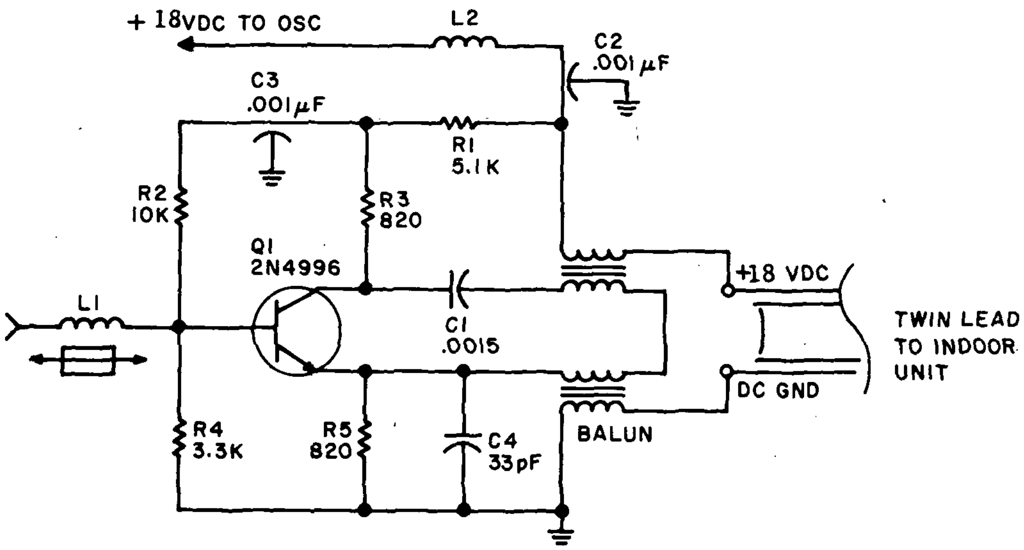


Figure 8. Type A Converter, IF Amplifier

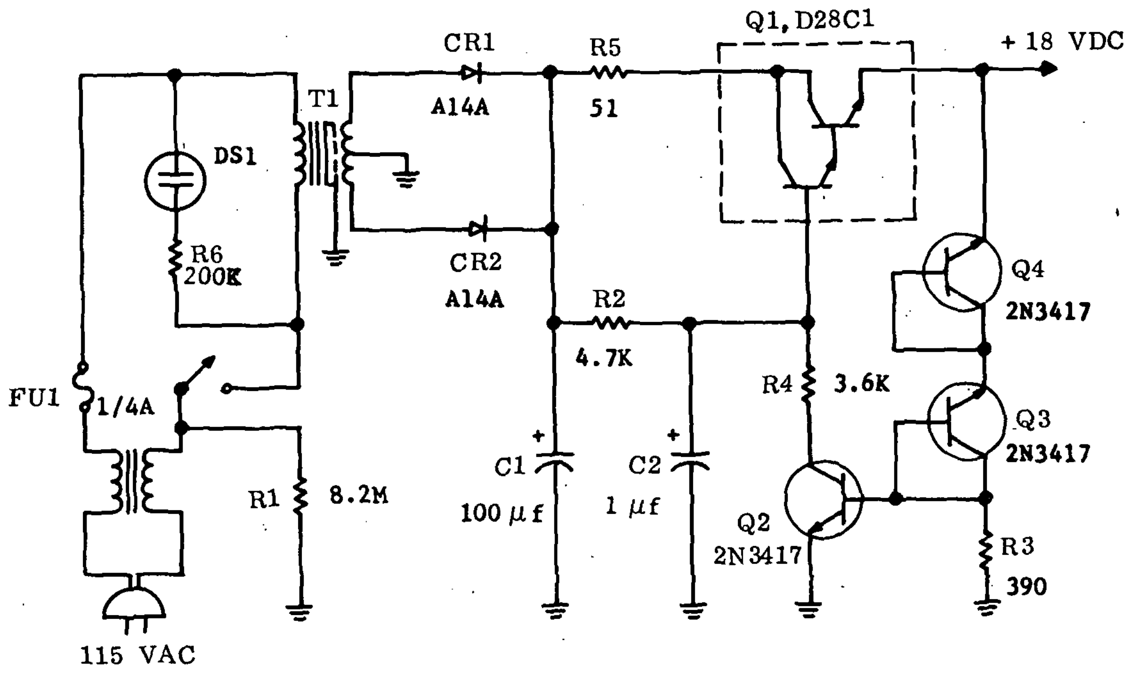


Figure 9. Type A Converter Power Supply

aluminum can for weather protection. A mounting bracket is welded to this cover to facilitate attachment to a mast or antenna assembly.

Four circuit board assemblies are used for the antenna unit functions. The largest circuit board consists of the double ground plane strip-line circuits used for the SRD resonant line section, the oscillator filter, and the mixer circuits. A small SRD circuit board is attached to the stripline assembly. The third printed circuit board contains the crystal oscillator and tripler circuits. The fourth board is for the IF amplifier circuit.

The large stripline assembly contains a backing plate with a right angle flange. This flange is bolted to a fitted base plate that fills the open end of the deep drawn cover. All input and output connections are made through the recessed baseplate. The oscillator-tripler circuit board and the IF amplifier board are attached to the backing plate with standoffs,

The printed circuit boards are layed out with holes spaced on a 0.1" grid to aid in production drilling. Flow soldering can be used for most component part assembly. Some component parts require hand soldering.

The indoor unit for the Type A converter contains the power supply and antenna switch. The power supply regulator circuits are constructed on a single-sided printed circuit board, which, together, with the power transformer, fuse, pilot light, etc., are mounted in a U-shaped sheet metal housing. The rear flange of the housing holds the antenna terminal strip and ac power socket. The front flange is used to mount the pilot light and switch. A conjugate, U-shaped cover completes the indoor unit housing. This cover has a bracket to which is attached the ac power cord plug.

3.2 TYPE B CONVERTER

The Type B converter system uses the same transmission frequency as the Type A system; however, frequency modulation is used instead of amplitude modulation. The use of frequency modulation changes several design tradeoffs. The first parameter change is the mixer and IF bandwidth. The bandwidth requirements are determined by the modulation index, maximum modulating frequency, and the receiving system local oscillator drift. A bandwidth of 30 MHz was determined to be adequate for the Type B converter.

Additional functions are required in the FM converter to derive a NTSC compatible AM signal from the FM signal. The most direct approach is to demodulate the FM signal and remodulate a VHF television channel carrier with the resulting baseband signal in a NTSC compatible format.

3.2.1 Mixer

The bandwidth of the mixer designed for the Type A system was more than adequate to handle the FM signal and therefore the same design was used* The SRD multiplier filter is not used with the Type B converter mixer because of a different approach in the generation of the local oscillator power. The stripline circuit is, therefore, considerably shorter for this converter type than for the Type A system.

---

*Drawing SK56157-D91-110 Sheet 2. See page 149.

3.2.2 Local Oscillator

The local oscillator requirements are different for the FM converters than for the AM converters. The tolerable drift of the local oscillator is greater because the IF feeds a FM demodulator instead of determining the input carrier frequency to the television receiver, as the case in the Type A converter system. Excessive drift is objectionable only because it requires additional noise bandwidth preceding the FM demodulator. The tolerable L. O. drift limits selected as a guideline in the Type B converter design was ± 1.5 MHz.

The crystal oscillator-multiplier approach was tried for the original Type B local oscillator source. The spurious AM signals discussed an connection with the Type A converter local oscillator led to its downfall as a local oscillator for the FM system. These spurious L. O. signals represent a fairly basic problem in wideband FM converters. It is generally desired, from a demodulator linearity viewpoint, to have the FM receiver IF near or above the frequency band where low cost crystals are available, i. e., < 100 MHz. The design of a multiplier chain under these circumstances requires very good suppression of lower order crystal oscillator harmonics at the latter points in a multiplier chain. Most solutions to this problem use high Q trap and/or bandpass filter circuits. These circuits become more susceptible to manufacturing tolerances and temperature effects, which make this approach to the solution undesirable where low cost objectives and wide temperature specification are combined. Fortunately there are more practical solutions, as discussed later under the Type B converter heading.

The tuning required to hold the energy at $f_{LO} \pm f_{crystal}$ down to a tolerable level could not be maintained over the antenna unit temperature range without complicating the circuitry. Alternate methods for implementing the Type B converter local oscillator appeared more promising. Two options were available; first, a direct transistor oscillator that was frequency stable over the operating temperature range specified, and second, a direct local oscillator with a frequency control network. The latter approach is more practical in an FM system than an AM system because the gain and discriminator functions required for the AFC are already present. One drawback to this approach for these converters was the physical separation of the local oscillator and discriminator portions in the antenna and indoor units respectively. Potential problems existed in stabilizing the AFC loop with the cable delay, and keeping power supply ripple from affecting the VCO would require large filter capacitors on the twinlead power connection.

A direct transistor oscillator was designed with temperature compensation that was compatible. This oscillator uses airline strips as the frequency determining elements. The materials chosen for fabrication provide an inherently stable circuit that requires a minimum of temperature compensation. The temperature compensation of the frequency is provided by sensing the ambient temperature with a thermistor that is a series element in the oscillator bias. Increasing ambient temperature tends to increase the dimensions of the airline strips causing the oscillator frequency to decrease. The thermistor resistance is reduced by the temperature increase

and the voltage drop across the thermistor is decreased. The latter effect is a result of the constant current bias network designed for the oscillator. As the voltage across the thermistor decreases, the V_{CE} of the oscillator transistor increases. The increased V_{CE} reduces the transistor effective output capacitance and causes the oscillator frequency to try to increase. The thermistor resistance versus temperature curve has a characteristic that is very nearly the complement of the oscillator frequency versus V_{CE} characteristic. The gain of the compensation is established by selecting the proper shunt resistor for the thermistor. The power supply is adjusted with a series potentiometer located in the indoor unit. This adjustment accounts for the differences in supply voltage and compensation network resistances. The voltage supply adjustment also makes it possible to tune the local oscillator remotely to a limited extent.

The airline oscillator equivalent circuit is shown in Figure 10. The two resonant elements are quarter-wavelength microstrip lines with air dielectric. These lines are coupled with adjustable capacitive plates at the line ends. A simplified sketch of the circuit layout is shown in Figure 11. The lines are mounted at right angles to each other to reduce circuit lead lengths. One line is driven by the transistor collector and the feedback signal to the transistor base is coupled from the other line. The base feedback signal is obtained by placing the half watt base bias resistor, R1, in the high field region beneath the undriven line. This placement induces sufficient signal in the resistor and resistor lead to cause circuit oscillation. The placement of this resistor is critical to the oscillator operation but can be easily reproduced from unit to unit. This type of coupling minimizes the resonant circuit loading and improves performance over that obtained using a discrete coupling capacitor. The base input impedance of the oscillator transistor is relatively high as a result of the unbypassed emitter resistor and the magnitude of the feedback coupling capacitance is less than one picofarad at the operating frequency.

Referring to Figure 12, the bias voltage for the base of the transistor determines the V_{CE} bias point. The I_E is established by the voltage drop across the emitter bias resistors, R_E. This voltage drop is held constant by using a zener connected transistor between the emitter bias point and the base through a choke. As the supply voltage at the emitter bias point is changed, the V_{CE} bias will change while the emitter current in the oscillator remains constant. This circuit connection permits the oscillator to have its output frequency changed by varying V_{CE} with a minimal change in output power, which is predominantly controlled by I_E. The change in V_{CE} is effected by changing the voltage drop across the shunted thermistor, R_S, as the temperature changes. The amount of voltage change with temperature depends on the thermistor characteristic and the relative magnitude of the thermistor resistance and the resistor shunting it. The choke in the bias lead presents a high impedance to the feedback signal that is fed to the base.

3.2.3 IF Amplifier

The center frequency for the IF amplifier was selected as 120 MHz, which is a tradeoff between two conflicting parameters, IF gain per stage and discriminator linearity. Available discriminator linearity is proportional to fractional bandwidth, thus the higher intermediate

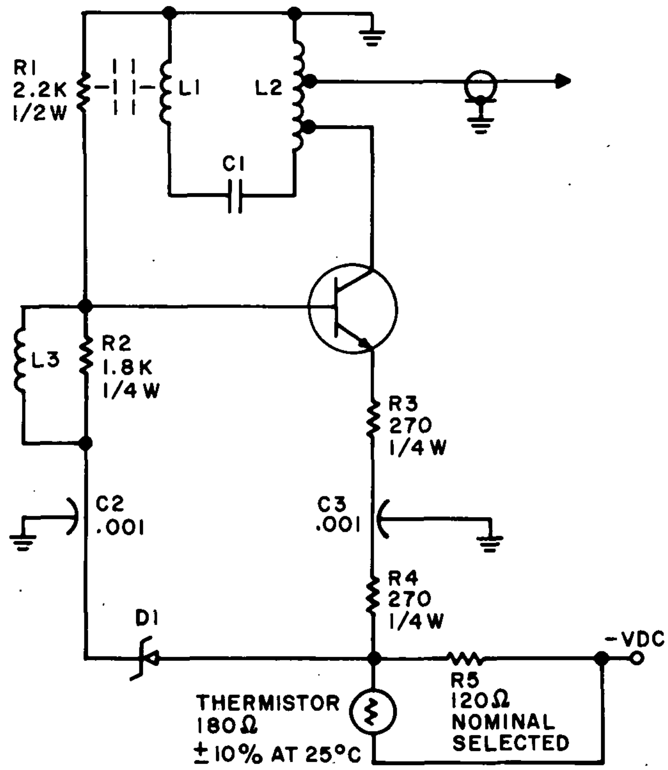


Figure 10. Type B Converter, Airline Local Oscillator Circuit Diagram

frequencies would ease this problem. The available gain per stage in the IF amplifier, as well as device noise figure, are improved at lower intermediate frequencies, especially when transistor cost is included as a factor. The IF amplifier uses 2N4996 transistors, which provide maximum gain and noise figure performance per unit cost at a frequency that permits the realization of adequate discriminator linearity.

Seven stages of IF gain were used in the design to reach the desired signal level at the limiter input with minimum antenna signal input. Three stages of gain are included in the antenna unit with the remainder in the indoor unit. This partitioning of the gain results in a signal level on the interconnecting twinlead that is not susceptible to interference. The partitioning also reduces crosstalk between the low level IF amplifier stages and the higher signal levels that exist prior to the discriminator.

All IF amplifier stages are essentially identical, a typical circuit is shown in Figure 13. The amplifier uses tuned transformer loads and inductive interstage coupling networks. The collector load is a bifilar transformer with the primary shunt tuned and the secondary resistively loaded.

The amplifier can be biased, either from a positive supply to ground or from ground to a negative supply. The bias voltage polarity depends on the requirements of the other circuits in the same package. A negative supply voltage is used in the Type B converter because the local oscillator requires a negative bias voltage.

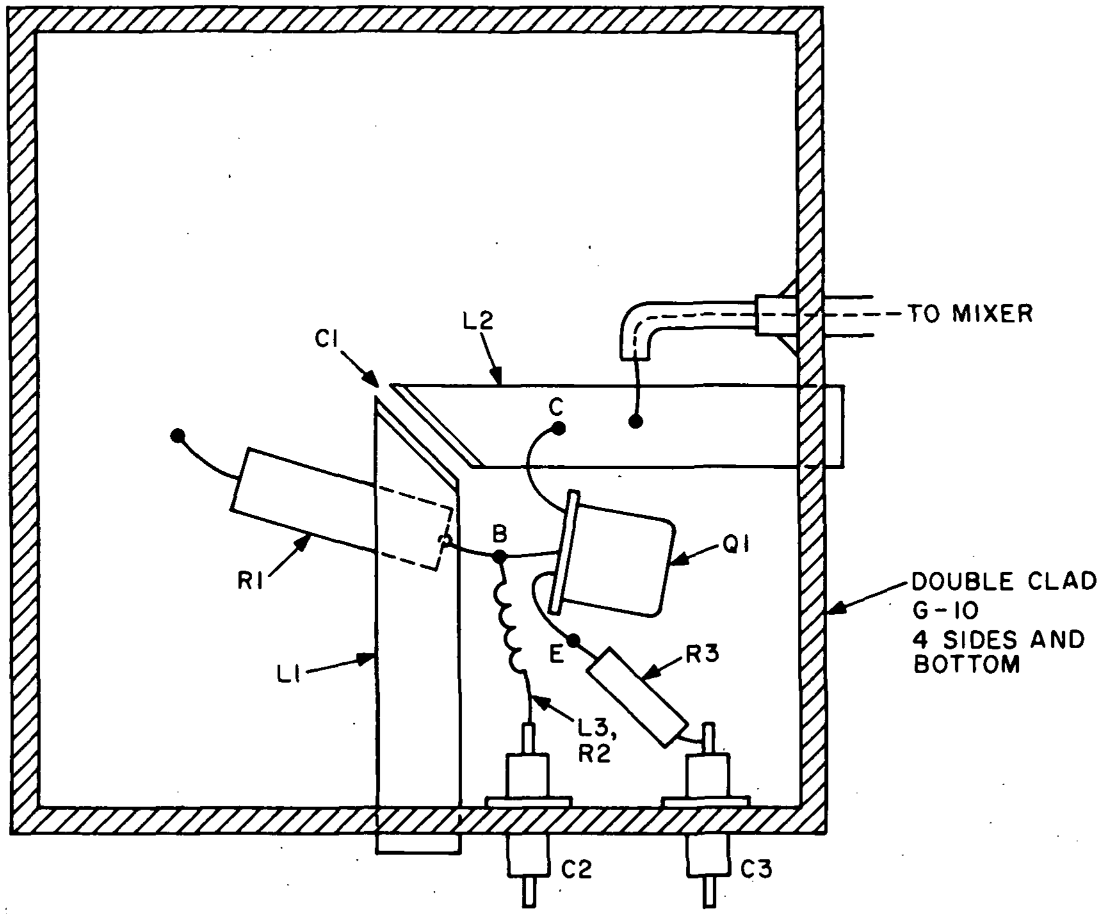


Figure 11. Type B Converter, Airline Local Oscillator Circuit Layout

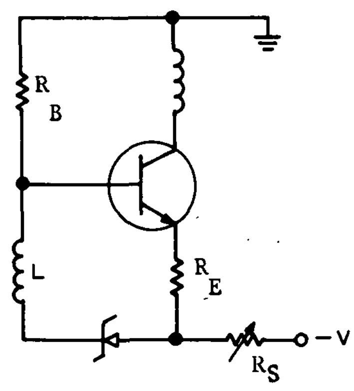


Figure 12. Type B Converter, Oscillator Bias Circuit

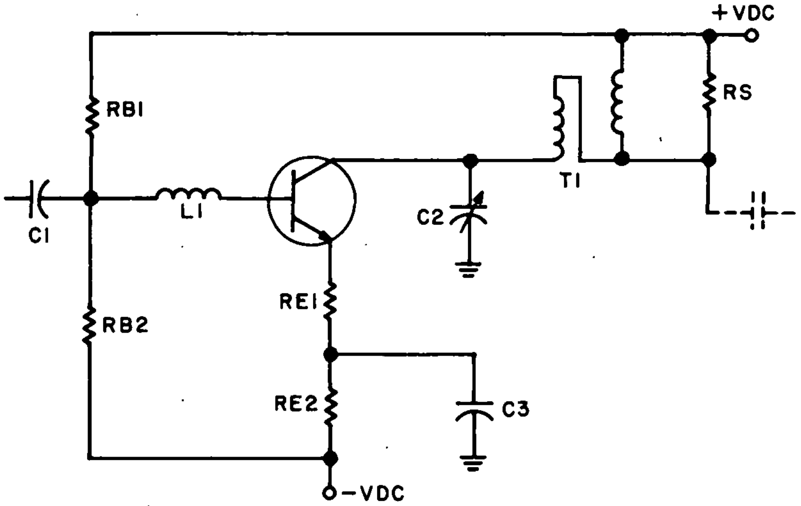


Figure 13. Typical FM Converter IF Amplifier Stage

The current level in succeeding stages is increased to accommodate the larger signal levels. The values of R_{B1} and R_{E2} are designed to set the desired current levels. The degenerate resistor, R_{E1}, is used only in the last three IF amplifier stages for stabilizing the circuits that operate at the highest current levels. The bifilar configuration introduces a 4:1 impedance step-down from the collector to the interstage network. The cold side of the secondary provides the collector dc bias path.

Balun connections are used at each twinlead termination to intercouple the two IF amplifier sections and to provide a dc bias path between the power supply and the antenna unit circuits. The balun ac connections provide a 75Ω amplifier termination and a 300Ω twinlead termination at each end of the interconnecting twinlead line.

The bandwidth of the IF amplifier is approximately 40 MHz and it can be narrowed slightly by appropriate alignment. The swept frequency and phase response of the IF is shown in Figure 14. The IF amplifier noise figure achieved was 4 dB ± 0.5 dB. The transistors used in fabricating the delivered units were screened and those exhibiting the best noise performance were assigned to the first and second IF amplifier stages. The screening procedure involved inserting the transistors as first IF units in a complete breadboard amplifier and measuring output noise power and amplifier gain. Those units exhibiting the maximum gain and minimum noise were reserved for use as mentioned above. This selection process

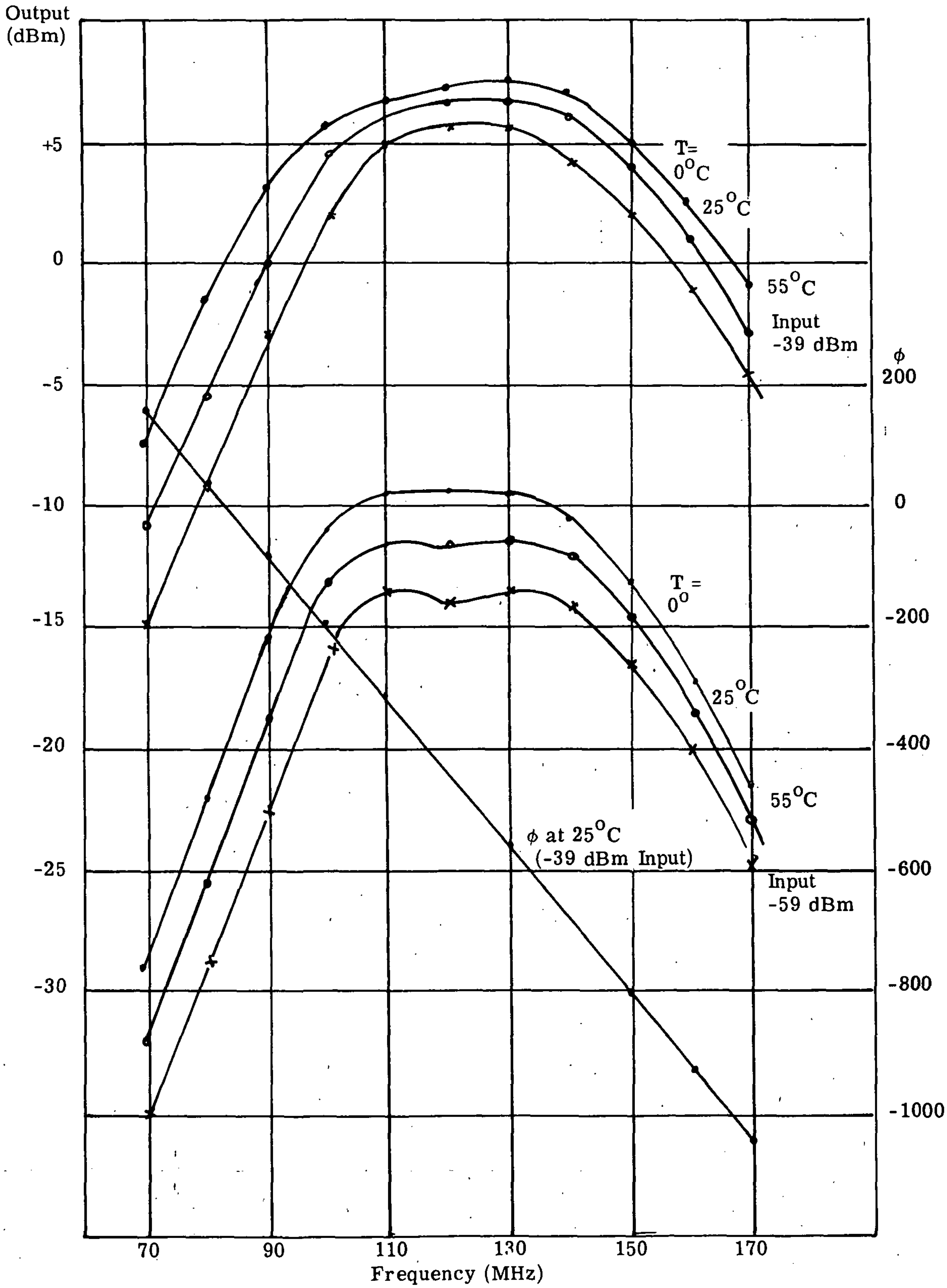


Figure 14. IF Amplifier Characteristics

represents one of the means available to manufacturers to increase the value of component parts procured in large quantities while minimizing cost.

Printed bifilar transformers and printed coils were used as the collector loads and interstage inductances throughout the IF amplifier. Collector tuning was done with low cost ceramic trimmer capacitors.

3.2.4 IF Limiter

The IF limiter, Figure 15, consists of a single stage differential amplifier. This circuit has the advantage that the input signal required to reach the limiting level is much less than that required for passive diode limiters. The differential amplifier also provides a convenient point for introducing a receiver output level control. The limiter output can be adjusted by controlling the current source magnitude, with R21, which controls the differential amplifier transconductance gain.

The total gain in the receiver system through the IF demodulator, video amplifier and remodulator is adjusted by varying the limiter current source magnitude. The tuned load of the limiter is returned to ground as are the bases of the differential amplifier. This bias arrangement uses the base-emitter voltage drop in the transistors to establish the V_{CE} operating point in the limiter. The 2N4996 transistors have excellent current gain at low values of V_{CE}. This permits the simple bias technique and eliminates offset and balancing problems in the limiter.

A filter is used between the limiter and discriminator to attenuate any harmonics of the instantaneous intermediate frequency generated in the limiter. The filter includes: the limiter tuned load and damping resistor, L_7, C_{18}, C_{21} and R_{24}; a series tuned section, L_8 and C_{22}; and a shunt parallel tuned section, L_9 and C_{23}. All inductors are printed as in the IF amplifier.

3.2.5 Discriminator

The filter drives a balanced discriminator, Figure 16 whose reactive elements are fabricated with one eighth-wavelength open and shorted transmission lines, which are also printed in microstripline form on the circuit board. The transmission lines are fed from a common filter output through individual resistors. These resistors, R25 and R26, match the characteristic line impedance of 120 ohms. A balanced differential video signal is obtained from the demodulator through two Schottky diode rectifiers.

A picture of the discriminator characteristic is shown in Figure 17. The schottky diodes used as detectors in the discriminator are the same type used in the S-band RF mixer. The balanced output is subsequently used to drive a differential video amplifier on another circuit board. The intermediate frequency components are filtered from the video outputs with a differential filter, the network parts of which are used to interconnect the printed circuit board. This filter also acts as a bias path for the video amplifier from ground through L_9, R_{26} and R_{25} and the demodulator diodes. The base current required by the video amplifier prebiases the demodulator

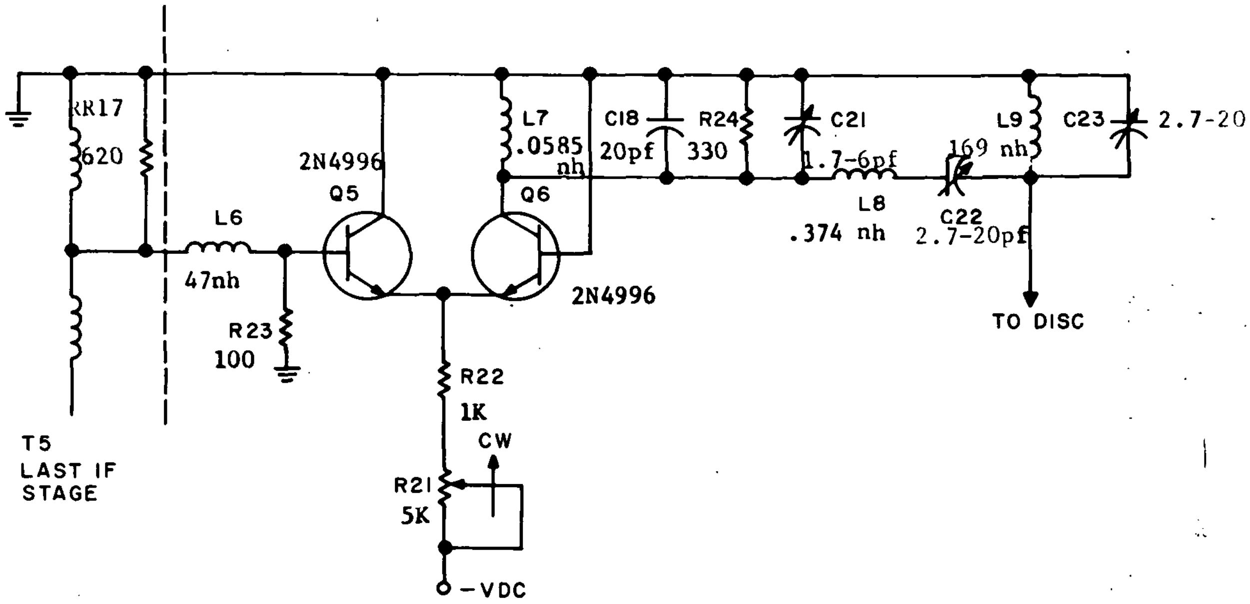


Figure 15. FM Converter, IF Limiter and Filter

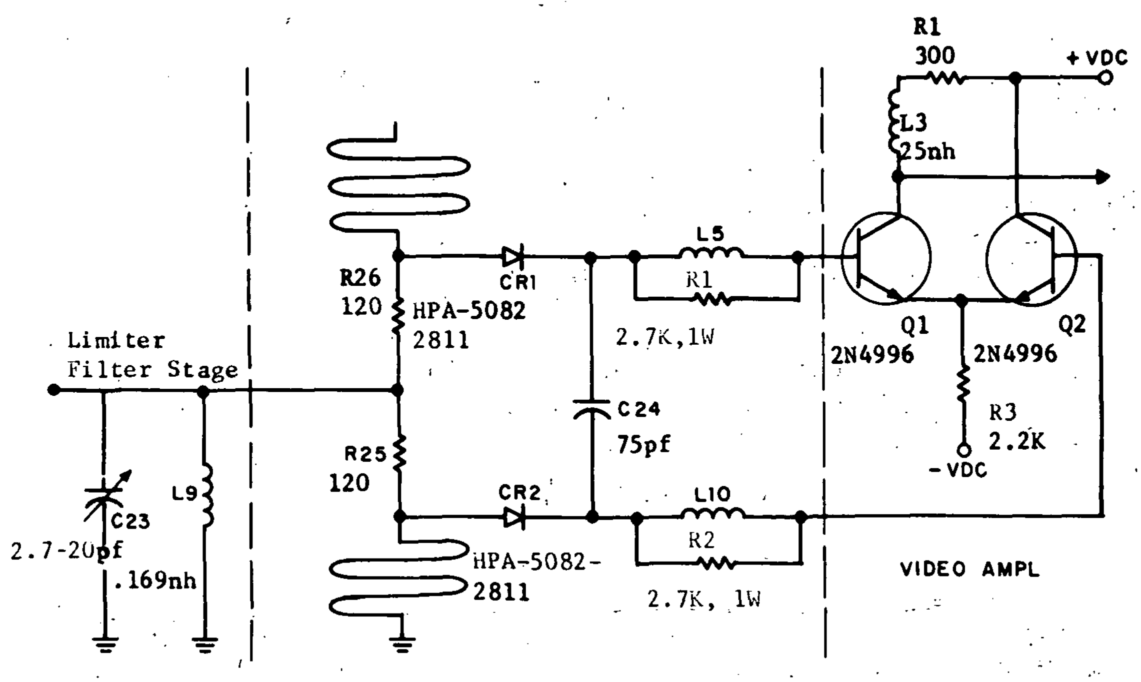


Figure 16. FM Converter Discriminator

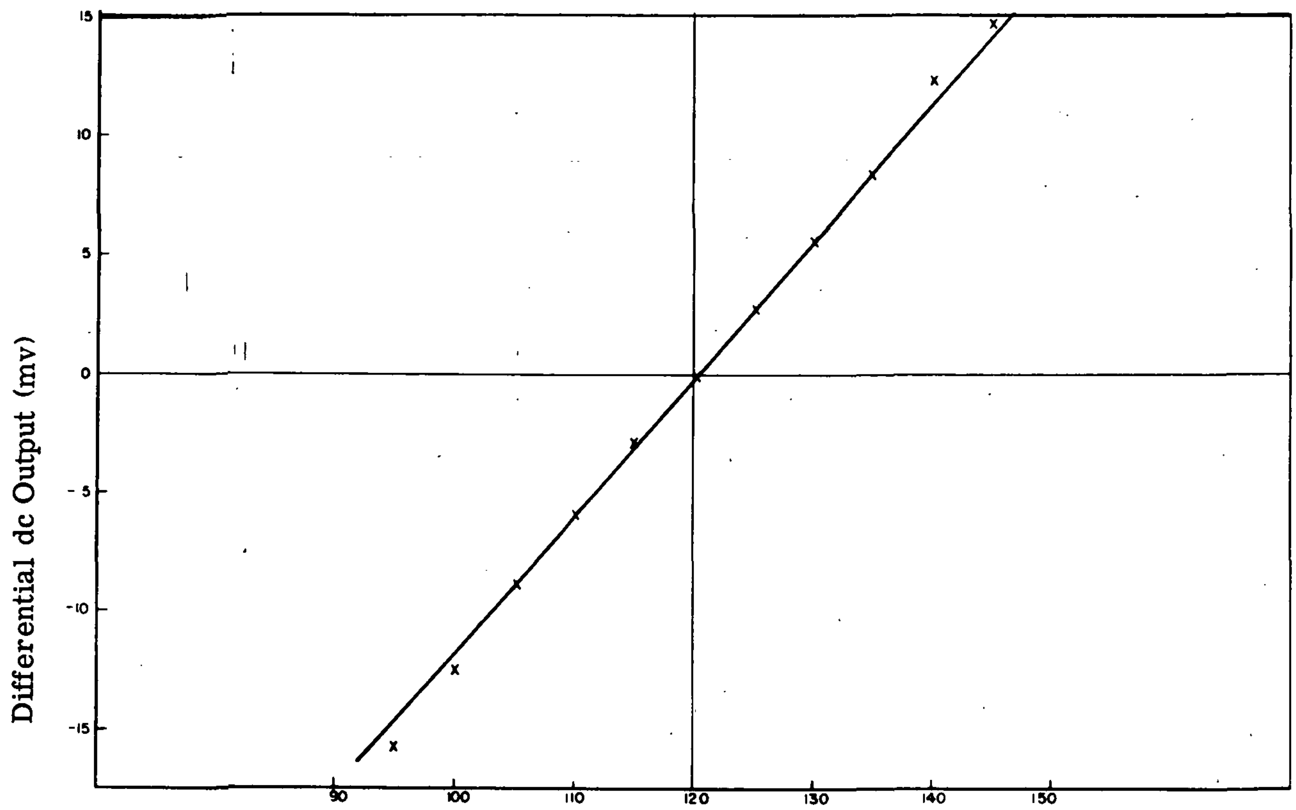


Figure 17. Discriminator Response Curve

diodes to a conducting state statically and, thereby, reduces the diode detection threshold.

The differential signal from the discriminator has greater linearity than could be obtained using a single transmission line discriminator at the same intermediate frequency. The discriminator outputs are filtered in the low pass filter consisting of L_5, L_{10} and C_{24}.

The inductors in this filter are used as board interconnections and consist of 24 turns of #26 enameled wire wound on 1 watt 3.3 KΩ resistors. The resistors act both as inexpensive coil forms and damping resistors in the filter.

3.2.6. Video Amplifier

The video amplifier, Figure 18 is biased from the positive nine volt supply for the collectors to the negative 18 volt supply for the emitter current source. A single sided video output with positive sync polarity is provided by the video amplifier. The bandpass of the filter and amplifier is sufficient to provide essentially flat gain beyond the audio subcarrier frequency of 4.5 MHz. The composite video and sound subcarrier signal is, therefore, preserved through the video amplifier. A peaking coil of 25 μh, L_3, is used in the video amplifier collector to shape the passband.

The composite video plus sound subcarrier signal is then dc restored through C_3 and CR_1 before it is fed to the remodulator circuit. The television receiver vestigial response characteristics eliminate

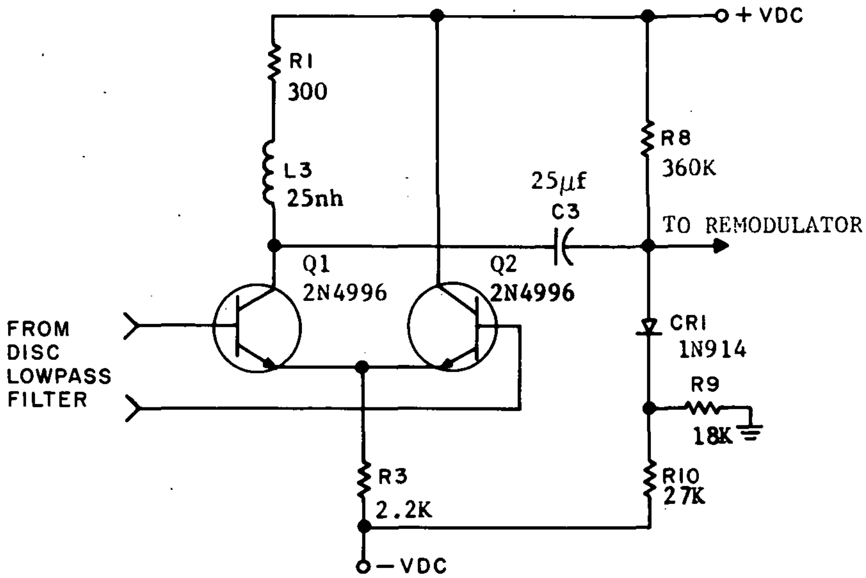


Figure 18. FM Converter Video Amplifier and DC Restorer

the lower sideband noise energy in the tuner and the upper sideband energy in the IF selectivity. The excess noise causes no discernible reduction in the television receiver video signal-to-noise ratio as long as the linear dynamic range of the converter remodulator and the television receiver circuits is not exceeded.

3.2.7. Remodulator

The remodulator circuit Figure 19, consists of a modulated current source and a transconductance multiplier. Q_4 is a modulated current source in which the collector current is controlled with the dc restored composite video-plus-audio subcarrier signal. The video signal is dc restored on the sync tip, which establishes the maximum collector current level from Q_4.

R_8 acts as a current source to bias the restorer diode, CR_1, to a level determined by R_9 and R_{10} from the negative supply. The current from R_8 also biases Q_4 on and sets the maximum Q_4 collector current level through R_{11} and R_{12}. The circuit is temperature stabilized by the diode drops of CR_1 and the base-to-emitter diode of Q_4. The negative going video components cause a reduction in the collector current of Q_4. The white modulation level is set by varying the converter receiver gain with the limiter current source control. The control compensates for the component gain variations in unit to unit from the IF amplifier through the AM remodulator. The video drive level is adjusted to obtain a 15% of sync peak carrier level at the remodulator output for a reference white video input. This adjustment will affect the total system output signal to noise ratio if the modulation index of the converter is low.

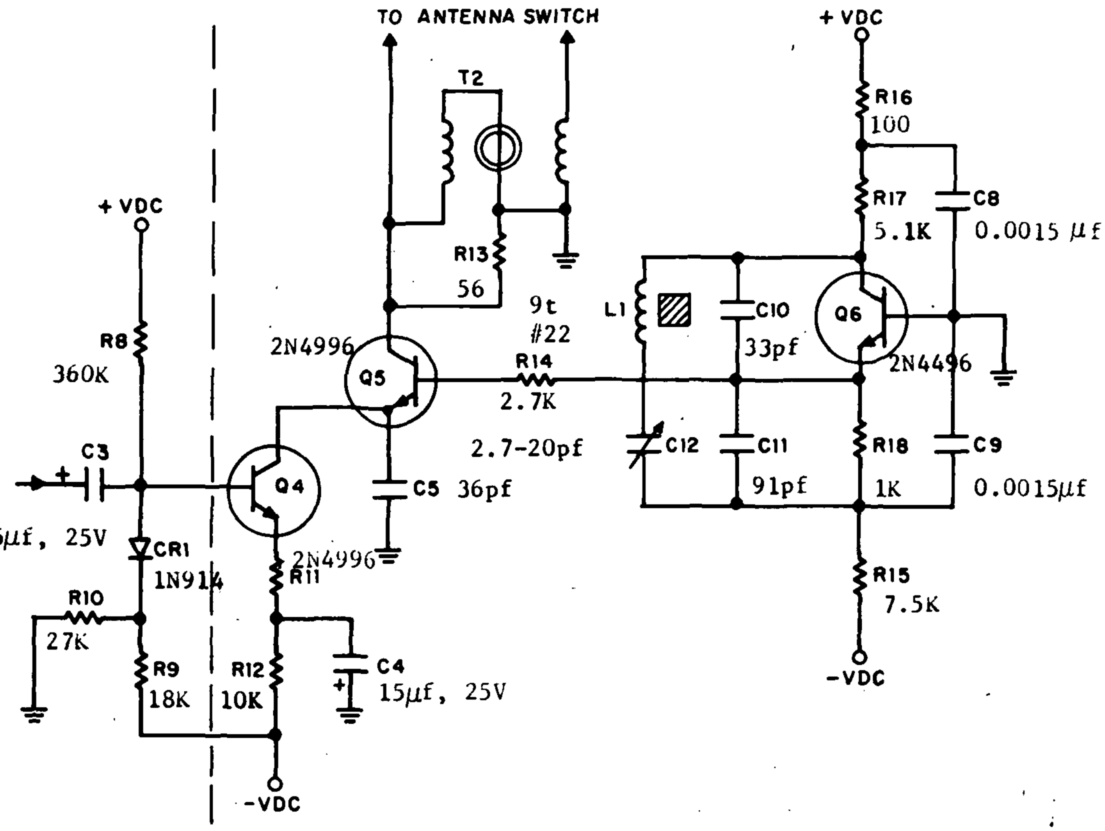


Figure 19. FM Converter Remodulator Circuit Diagram

The collector current from Q_4 sets the average emitter current in Q_5 and thereby establishes the level of carrier frequency (injected from the oscillator to the base of Q_5) that appears as collector current. C_5 acts as an emitter bypass at the carrier frequency. The modulated, carrier output current at the collector of Q_5 drives a balun. The primary is shunted with 75Ω, R_{13}, to establish the source impedance of the remodulated output. The balun matches the 75Ω source impedance to the 300Ω twinlead, which carries the remodulated signal to the television receiver through the antenna switch.

The local oscillator, used to provide the channel carrier signal, which is remodulated, is a standard Colpitts oscillator with the output taken from the emitter feedback node. This provides a convenient and stable bias voltage for the remodulator base,which permits the balun in the remodulator collector to be returned to ground. The oscillator base is also grounded and the oscillator bias current is set with emitter resistors, R_{18} and R_{15}, from the regulated negative supply. The oscillator frequency is adjusted with a variable capacitor, C_{12}. The oscillator is decoupled from the positive collector bias supply by R_{16} and C_8. R_{15}, the large emitter biasing resistor, is bypassed with C_9 to set the appropriate emitter impedance.

3.2.8 Power Supply

The power supply, Figure 20 used in the Type B converter provides two output voltages. The negative 14.5 V supply is the critical supply in the system and is, therefore, regulated with an active feedback regulator. The regulator circuit provides the dc voltage control and ripple reduction required at a lower cost than Zener and capacitor regulator-filter circuits. The circuit used is a shunt regulator with a power Darlington device as the current bypass element. The shunt element is turned on when the reference diodes in the base are forward biased. The high input impedance of the Darlington assures that the Zener current variation is small; hence, the Zener voltage is not modulated by the ripple at the sampling point.

The reference diodes used in the base circuit of the Darlington are low cost transistors with selected base emitter breakdown voltages and Zener characteristics. The collector to base connection temperature stabilizes the Zener characteristic and increases the power handling capability of the devices.

The positive supply is used as the non-critical collector bias supply for the video amplifier, modulator current source and oscillator collector supply functions. This voltage is regulated to a convenient level of about +9V dc with a single Zener connected transistor and dropping resistor, Q_9 and R_{22}. The unregulated dc for the negative and positive power supplies is generated from two full-wave rectifiers with capacitive filters. The rectifiers are driven with a center tapped, step-down transformer. The transformer sets a convenient ac voltage and decouples the circuits from the line voltage. The transformer primary is fused and RF isolated from the line with a twin choke, L_2, to prevent converter-to-line RF coupling.

The converter uses a three-pole double-throw switch to switch converter output signal sources and to turn the ac power to the converter on and off. Two switch sections are used to transfer the output twinlead from a local antenna source to the converter remodulator output when the converter is turned on. The other terminals switch the ac power to the transformer primary, a neon lamp, connected across the transformer primary, indicates when the ac power is applied to the converter. A low cost slide switch was selected for the converter when tests indicated that the RF attenuation and feedthrough characteristics were minimal. The power switching arrangement eliminates any converter interference when not in use.

3.2.9 Type B Converter Packaging

The S-band FM converter is packaged in two chassis. The antenna unit components are mounted on a common plate that supports the oscillator, mixer and IF preamplifier. A base plate is perpendicularly attached to the mounting plate and the entire assembly is housed in a deep drawn aluminum cover. The cover has a plate welded to the outside to facilitate mast attachment.

The base plate mounts the RF and twinlead connections, which are recessed in the bottom of the antenna unit. The configuration provides easy access and rain protection. External parts are anodized.

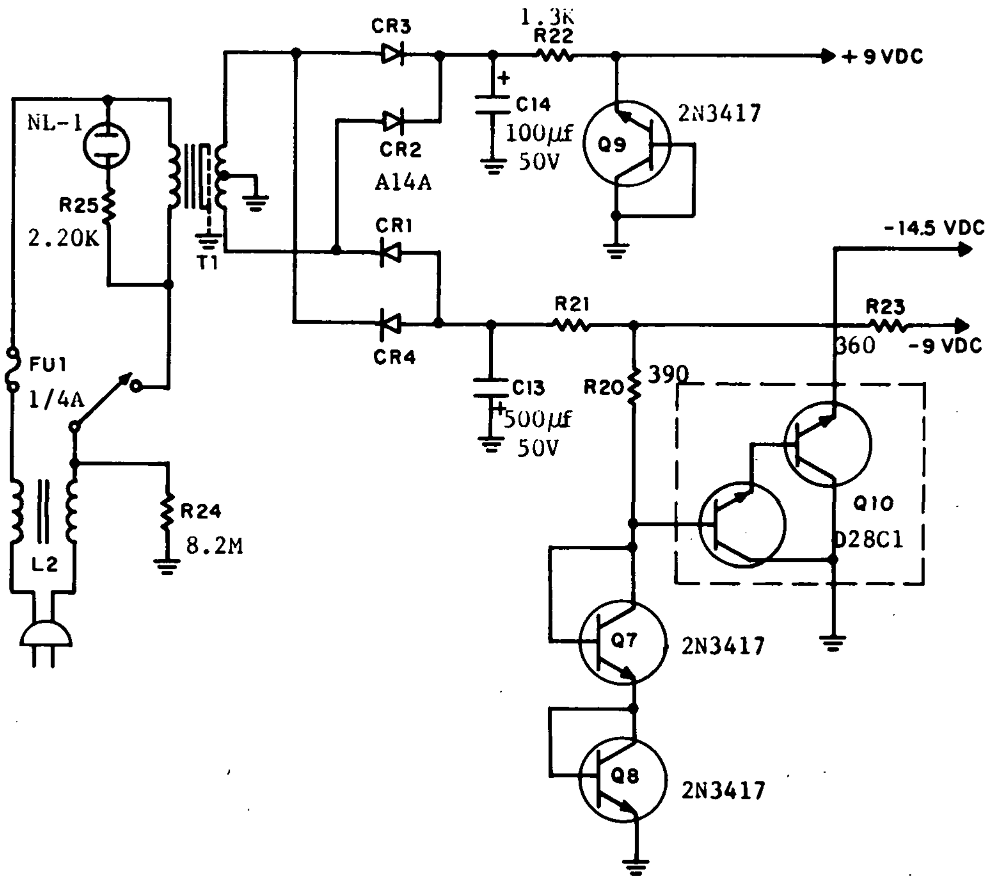


Figure 20. Type B Converter Power Supply Circuit Diagram

The indoor circuits are mounted in a U-shaped chassis. The switch and pilot light are mounted on the front. The power connection socket and twinlead terminal strip are mounted on the rear. The power transformer, fuse block, and circuit boards are bolted in the chassis with the IF-Limiter-Discriminator board stacked on top of the Power Supply - Video Amplifier - Remodulator board. A conjugate U-shaped cover attaches to the indoor chassis.

The fabrication techniques used to package the converters do not represent the optimum approaches that would be followed in mass production. The engineering models are, therefore, not representative of the packaging considered in the cost analysis. Prohibitive tooling costs prevented the use of plastic chassis parts or die cast cavities as would likely be used in production. All other assemblies within the converters are close to typical production procedures. The exceptions are; hand soldering instead of flow soldering, and the use of nuts and bolts instead of rivets. Some hand assembly procedures (such as the local oscillators in the S-Band Converters) require additional product design prior to production release.

3.3 TYPE D CONVERTER

The Type D converter uses an FM transmission format as in the Type B converter; however there are two differences, the transmission frequency (12.0 GHz versus 2.25 GHz) and the modulation index (3.0 versus 2.0). The larger modulation index in the Type D converter requires a predetection bandwidth of about 39 MHz including an allowance of 3 MHz for local oscillator drift. The same IF of 120 MHz is used in both FM converters. The FM transmission is demodulated to obtain the baseband composite video signal plus aural subcarrier as in the other FM converter. Because of the similarity in functional requirements, the same circuitry is used for the IF amplifier, limiter, discriminator and subsequent AM remodulation in each FM converter.

The other system parameter change, the transmission frequency, results in the differences between the two FM converter designs. The X-band frequency allocation requires a different approach in the design of the front end circuits.

3.3.1 Mixer

The mixer circuit selected for the X-band FM converter is a balanced mixer fabricated in stripline format.* High frequency, beam-leaded, Schottky diodes are used to minimize the circuit parasitic reactances. The RF port of the mixer is designed to interface directly with the RF coaxial connector. The local oscillator port uses a short capacitive probe to extract energy from the local oscillator.

3.3.2 Local Oscillator

The original approach for generating local oscillator power was the use of a crystal oscillator - multiplier chain. This approach was discarded for the reasons discussed in section 3.2.2 and because of the difficulty in obtaining a temperature stable and efficient multiplier circuit using a step recovery diode. A cavity type oscillator using a Gunn diode was chosen instead. The advantage of the Gunn oscillator lies in its ability to provide direct dc to microwave power conversion. A potential problem was the frequency stability with temperature. This problem was solved by incorporating a temperature compensating structure into the cavity. The basic cause of frequency drift with temperature is the dimensional change of the cavity structure with temperature.

One possible solution is to fabricate the cavity with temperature stable material such as Invar. This solution is expensive because it is expensive to machine temperature-stable metals. The use of a temperature compensating structure is less costly and provides good performance.

The cavity structures used for Gunn oscillators frequently use a dielectric probe for frequency adjustment. A two-dimensional sketch of such a cavity structure is shown in Figure 21. The cavity structure and the Gunn diode are both relatively broad band and the frequency can be adjusted with dielectric loading within the cavity. When the dielectric material, usually quartz or alumina, is inserted further into the cavity the resonant frequency

*Drawing Sk56157-D91-37. Sheet 2. See page 150.

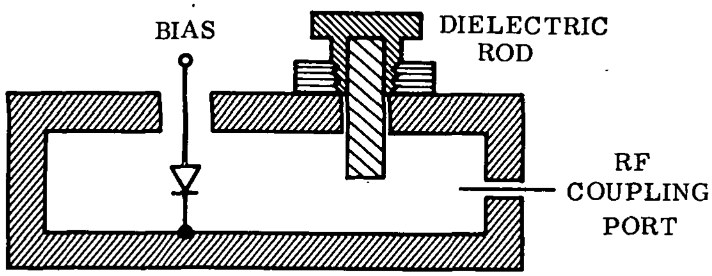


Figure 21. Simplified Gunn Oscillator Cavity Structure

of the cavity is lowered. This technique is modified slightly as in Figure 22 to provide temperature compensation. In this case an alumina rod was selected as the tuning and compensating element. The thermal coefficient of expansion for the alumina rod is much smaller than the aluminum cavity. The tuning rod is suspended or held at a distance, h, from the cavity wall. As the temperature increases the cavity, together with the supporting column for the tuning rod, will expand. The incremental increase in the distance, h, can be predicted. The rod length, l, from the point of suspension will also increase but by a much smaller amount. The net effect of a temperature increase is to cause an extraction of the tuning rod from the cavity by an amount proportional to the differences in the thermal expansion coefficients of aluminum and alumina. This ratio determines the set point for rod suspension that will give the best compensation. When properly constructed the thermal increase in cavity size, which would cause a frequency reduction, is balanced by the thermal extraction of the tuning rod, which would cause a frequency increase.

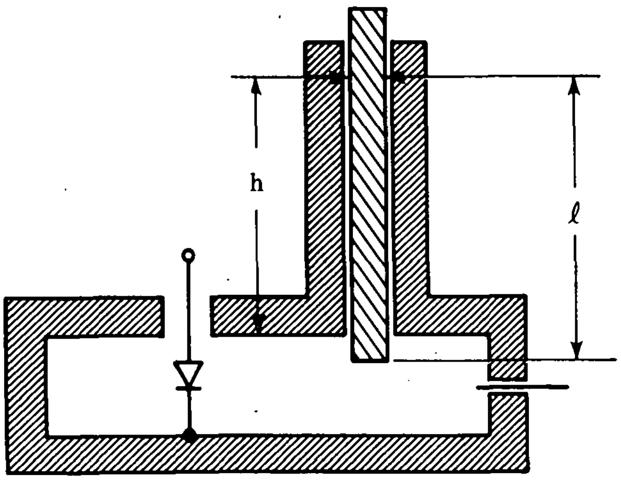


Figure 22. Simplified Gunn Oscillator Cavity with Temperature Compensating Tuning Probe

This design was used successfully in the fabrication of temperature stable local oscillators for the Type D prototype converters. Additional product design refinements are warranted in this application to improve the ease of tuning. Figure 23 shows the temperature compensated tuning characteristic of the cavity design. A temperature stability of $\pm$ 1.15 MHz is possible with this approach at 12 GHz over an ambient temperature range of $-40^{\circ}C$ to $+55^{\circ}C$. The local oscillator power level at the mixer is adjusted by moving the pickup probe attached to the mixer into or out of the cavity structure.

3.3.3 IF Amplifier

The same IF amplifier design is used for the Type D converter as for the Type B converter. Additional IF bandwidth is obtained by altering the alignment procedure. The slight reduction in IF gain is compensated by the increased signal power in the Type D system and the greater FM improvement obtained from the wider deviation used.

The antenna unit IF amplifier is biased from a positive supply rather than a negative supply as in the Type B converter. This option is used to make the amplifier compatible with the local oscillator bias requirements.

3.3.4 IF Limiter-Discriminator, Video Amplifier and Remodulator Circuits

The remaining circuits in the signal path of the Type D converter are identical to those used in the Type B converter. The discriminator alignment is adjusted to accommodate the additional bandwidth required. The current source magnitude in the limiter is again used to set the receiver gain and the remodulation carrier level at white.

3.3.5 Power Supply

The power supply used in the Type D converter consists of positive and negative full wave rectified circuits with zener regulation in the indoor unit. In addition, a transistor regulator is used in the antenna unit to bias the Gunn diode and IF preamplifier circuits. The circuit diagram for the Type D converter indoor unit power supply is shown in Figure 24.

The primary side of the power transformer is fused and decoupled with the RFI choke, L_1. Line power is switched to the primary with a section of the antenna/power switch and a neon pilot light is provided as shown. The positive full-wave rectifier provides the partially filtered dc at 20V. The two watt resistor, R_{22}, protects the rectifier circuit from external shorts and relieves some power dissipation in the antenna unit regulator. This positive power supply must provide up to 600 ma of starting current for the Gunn oscillator with sufficient regulation to keep the transistor regulator in the antenna unit conducting. A positive 9 volts for the indoor unit circuits is also derived from this supply with the zener connected transistor Q_9 through R_{21} and R_2.

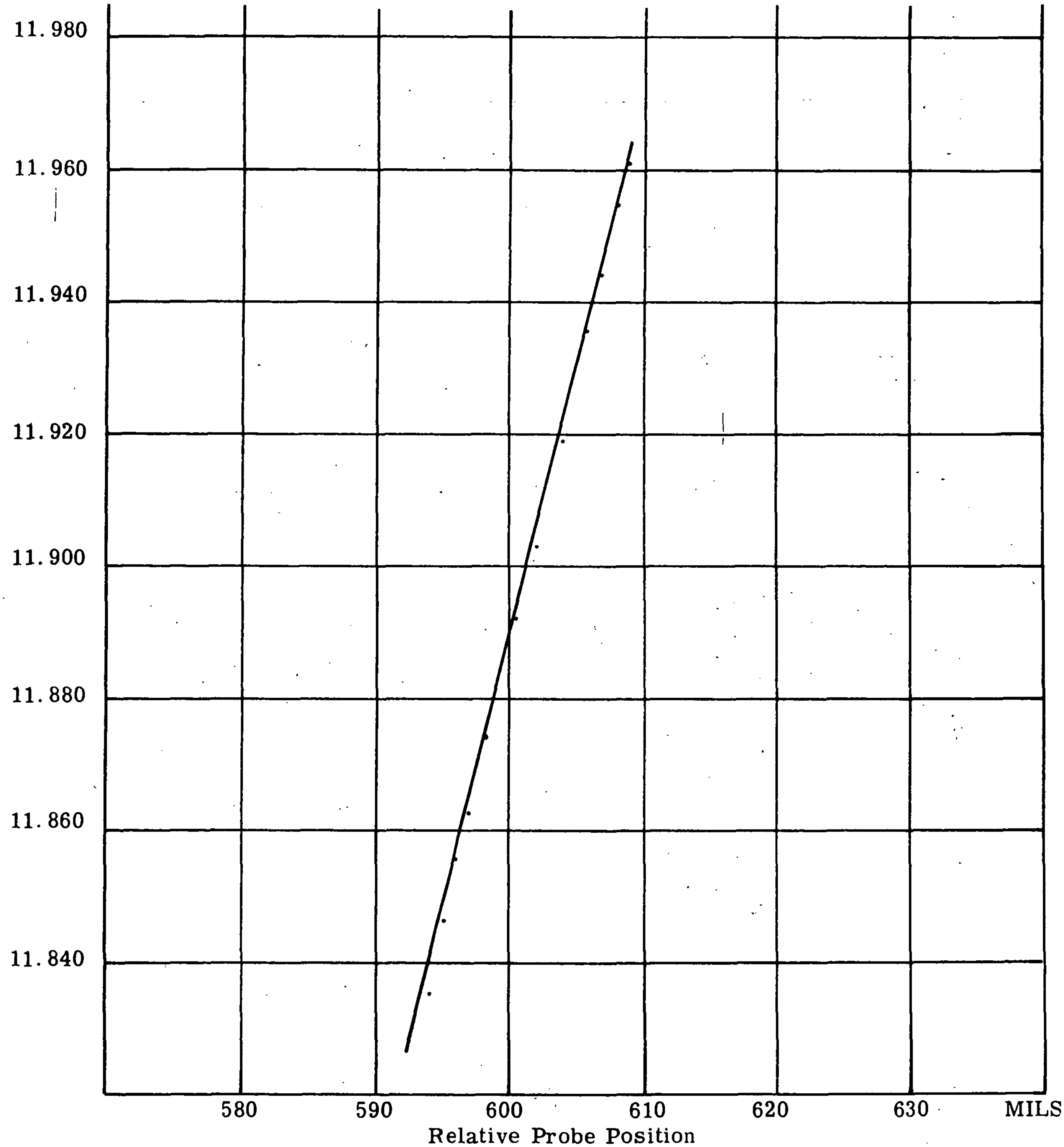


Figure 23. 12 GHz Oscillator Frequency Versus Probe Position in Vicinity of Operating Frequency

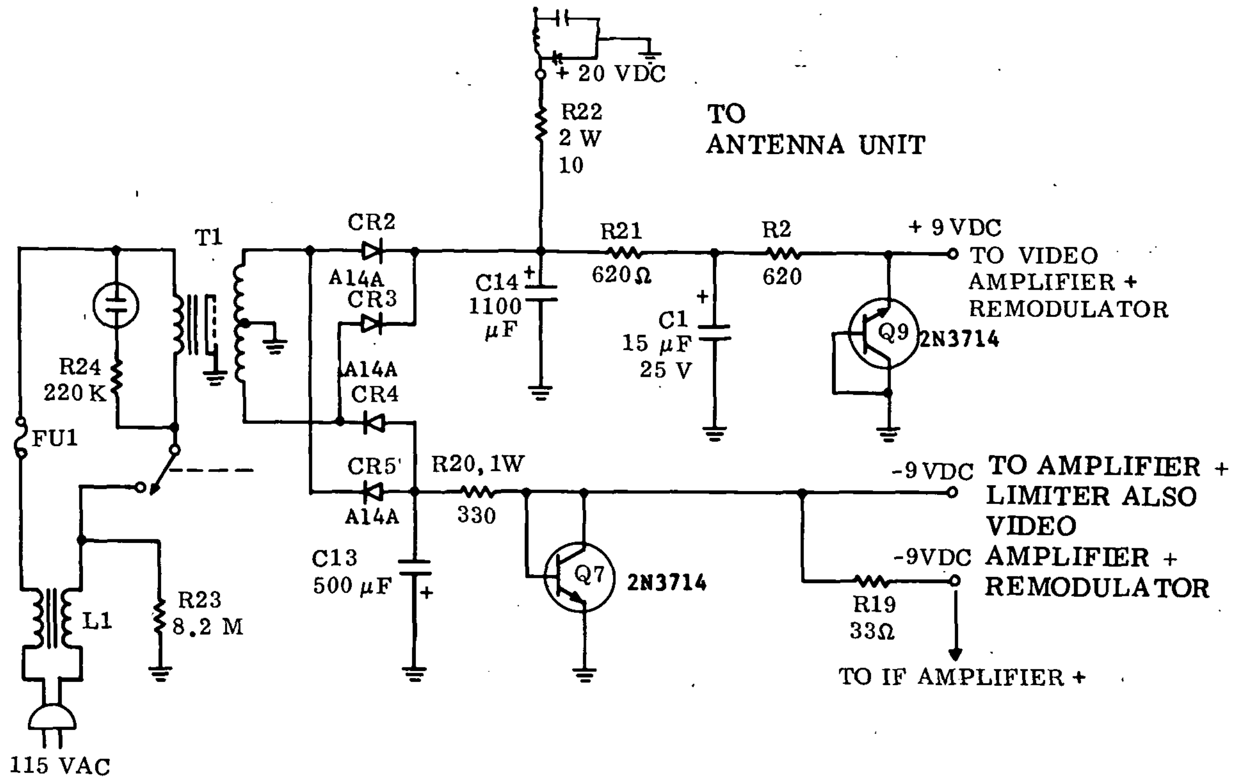


Figure 24. Type D Converter Indoor Unit
Power Supply Schematic

The negative, full-wave rectified supply is zener regulated using Q_7 through R_{20}. This supply provides the other bias requirement within the indoor unit.

The antenna unit regulator circuit is shown in Figure 25. The +20 VDC bias is fed to this regulator through the twinlead interconnecting cable from the indoor unit power supply.

Q_1 acts as a series regulator. The output voltage is sensed through the zener connected transistor, Q_2. Any excess output voltage causes the drive current to the series regulating transistor to be diminished. C_1 stabilizes the feedback loop in the regulator. C_2 suppresses potential oscillations with the Darlington connection by providing a low load impedance at high frequencies. Resistor, R_1, provides current limiting and minimizes power dissipation in the series regulator. This resistor can be removed if the Gunn diode requires higher than normal peak starting currents. The choke, L_1, isolates the oscillator and IF amplifier at RF.

The low output impedance provided by this regulator is necessary to operate the Gunn oscillator. The Gunn devices exhibit relatively small negative resistance characteristics and oscillation will not occur if there is an appreciable dc source impedance from the power supply regulator. Special care must also be taken to maintain a minimum of resistance in the Gunn diode regulated bias circuit. Gold-plated bias probes and diode holders are suggested to minimize contact resistance.

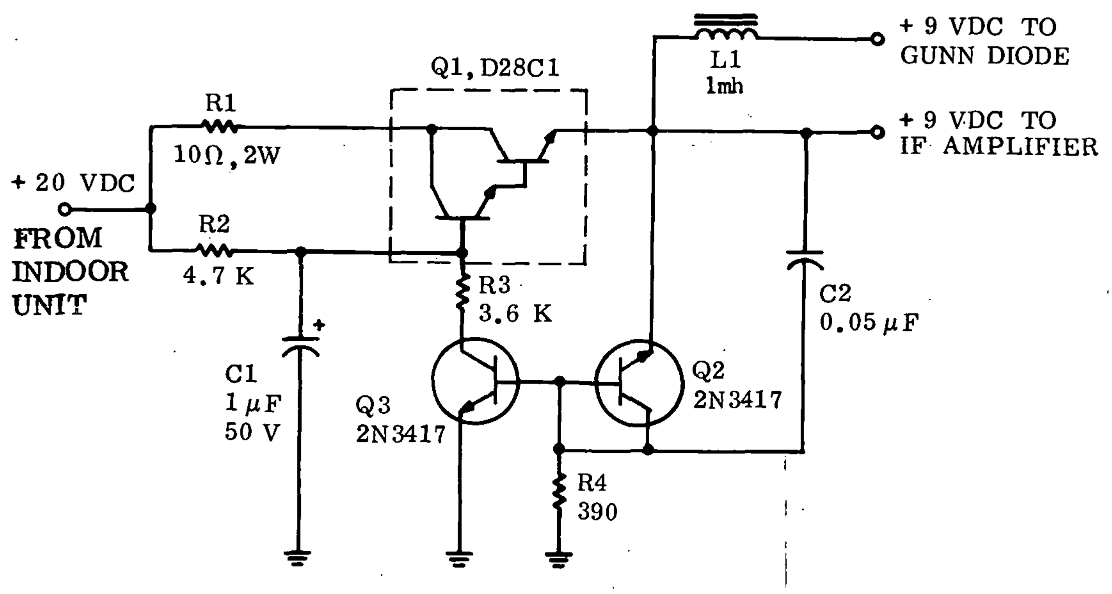


Figure 25. Type D Converter Antenna Unit Power Supply Regulator Schematic

3.3.6 Type D Converter Packaging

The prototype packages developed for the Type D converter antenna and ground units are very similar to those used for the Type B converter. The antenna unit is housed in a deep-drawn aluminum enclosure with a recessed base plate through which connections are made. The internal construction consists of a right-angle mounting bracket attached to the base plate. The subassemblies are attached to the mounting bracket, either directly or with standoffs, where required.[1] A major subassembly consists of the oscillator cavity on which the mixer is mounted. The prototype cavity is machined from aluminum; however, production quantities would be provided with castings. The Gunn diode is mounted within the cavity by a double-threaded diode holder.[2] The Gunn diode is screwed firmly into the diode holder using a thermal conducting grease for best heat dissipation. The diode and holder is then screwed into the cavity as a unit. The diode bias contact is made by forcing the diode against a bias probe that has been mounted in axial alignment through the opposite side of the cavity. This probe is plated on the end with gold and has a teflon insulation applied where it penetrates the cavity. The probe is potted in place with a resilient RTV potting compound, which provides the necessary support and insulation. The resilience of the potting compound maintains a relatively uniform contact pressure as the diode and diode holder assembly is forced against it. The resilience also allows for fine relative motion of the diode with respect to the cavity as a result of thermal expansion effects.

---

(1) See drawing SK56157-E91-104, 120 GHz FM Antenna Unit Assembly. See page 151.

(2) See drawing SK56157-C91-96, Mixer-Cavity Assembly. See page 152.

The cavity tuning probe and temperature compensating device is inserted into the cavity through the same wall as the diode bias probe. Opposite the tuning rod, a fourth hole provides entrance for the capacitive power probe that drives the mixer. The complete cavity structure has an irridite finish to maintain maximum electrical conductivity at the surfaces. The cavity is closed with a rectangular cover plate and six screws.

The tuning probe is fixed at the optimum point for temperature compensation with set screws. Fine tuning of the oscillator frequency is made by adjusting the probe position. The oscillator frequency is lowered as the probe is inserted into the cavity.

The cavity structure also provides support for the mixer assembly, which is constructed in stripline form with backing plates. The RF connector is soldered to the mixer and aligns with a hole in the baseplate of the antenna unit enclosure.

The IF amplifier board is mounted on the same side of the main bracket. This board is positioned away from the oscillator-mixer assembly with standoffs. The voltage regulator board is mounted on the opposite side of the bracket. A dc insulated heat sink and a standoff are used to hold this board in place.

The indoor unit enclosure for the Type D converter is constructed in the same manner as the indoor unit for the Type B converter. The only distinctions between the Type B and Type D indoor units are details in the power supply. A slightly larger power transformer is used in the Type D converter and a transistor regulator is not used in the Type D converter ground unit power supply.

4.0 TARGET COST CONVERTER

The "Target Cost" converter program was added to the scope of the original program to explore potential product cost savings that might result from reversing performance and cost emphasis. The main development program associated with the contract was oriented toward providing specific performance goals at minimum cost. The emphasis of the Target-Cost program was to determine the level of performance that could be achieved with a Type D converter system within a fixed cost restraint and with relaxed system specifications. Two main design constraints that were imposed on the previous development, and were relaxed for this study, were: (1) the available signal power level (2) the restriction on combining the TV receiver and converter functions at other than an RF interface.

The antenna output signal level for the target cost converter was raised from -105 dBW to -100 dBW. The relaxation of the TV receiver interface restrictions permitted the sharing of enclosures and power supplies, and the elimination of the AM remodulation function within the converter.

The target-cost converter program was organized as three tasks. The first (Task X) dealt with conceptual designs and preliminary evaluation of techniques; the second included the main development of a target cost converter; and the final task covered the fabrication, test, and evaluation of two target cost units.

4.1 DESIGN ALTERNATIVES CONSIDERED IN TASK X

The first design alternative used a straight-forward combination of converter and TV receiver components plus a simplification of the converter front end mixer. The block diagrams of the original and "target cost" 12 GHz converters are compared in Figures 26 and 27.

The dashed blocks in Figure 27 represent the television receiver's circuitry and the solid blocks represent the converter circuit. The converter of Figure 26 is a completely self-contained unit (including cabinet and power supply) that converts the 12 GHz F. M. signal received by the antenna into an amplitude modulated picture carrier and the frequency modulated sound carrier at VHF channel 6. In contrast to this the converter shown in Figure 27 converts the 12 GHz FM signal to an intermediate frequency and after appropriate amplification and limiting extracts the composite video and sound subcarrier. These signals are injected into the video amplifier and sound IF amplifier of the T. V. receiver. The indoor unit of the converter is housed within the T. V. receiver's cabinet and the converter's power is obtained from the receiver's power supply.

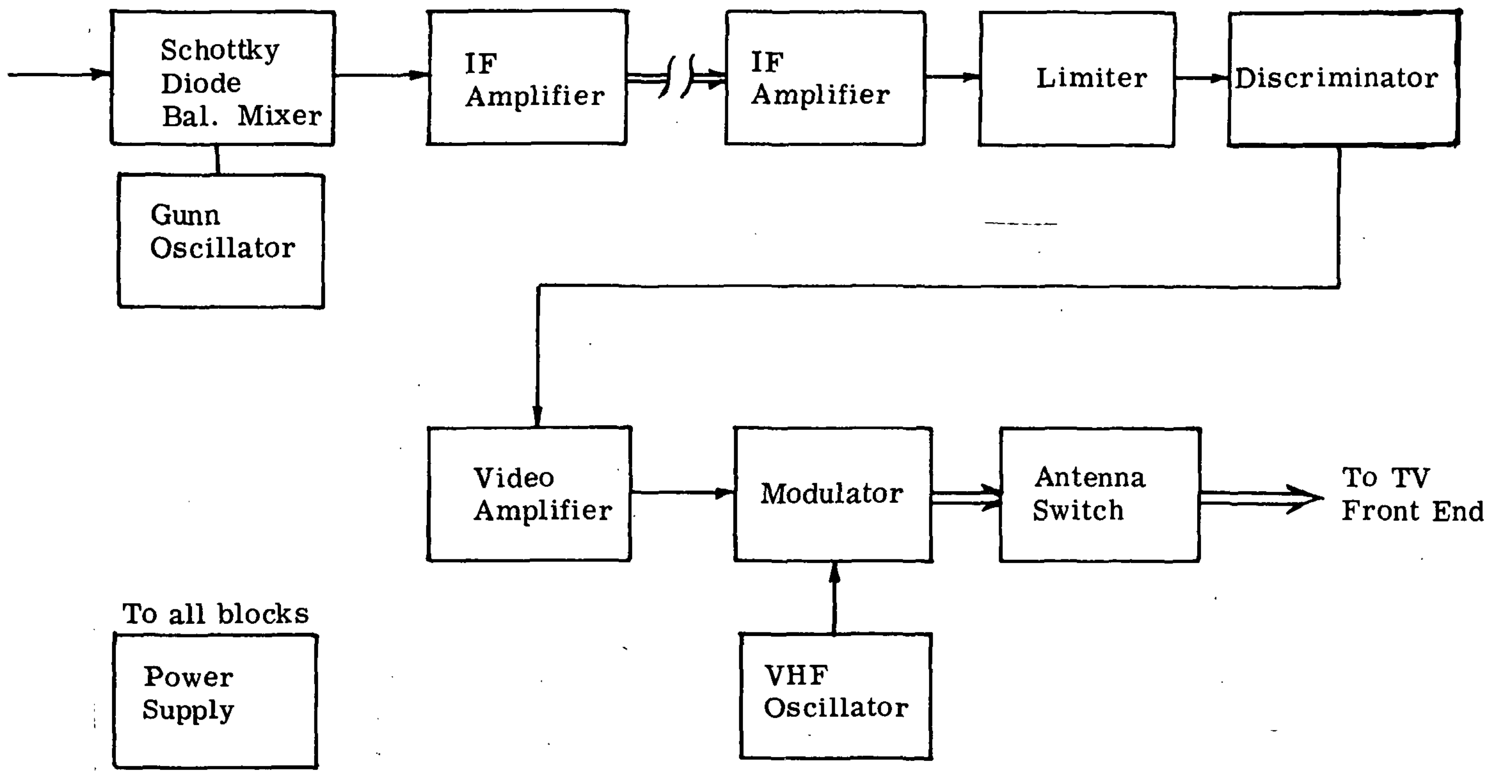


Figure 26. 12.0 GHz Converter Block Diagram

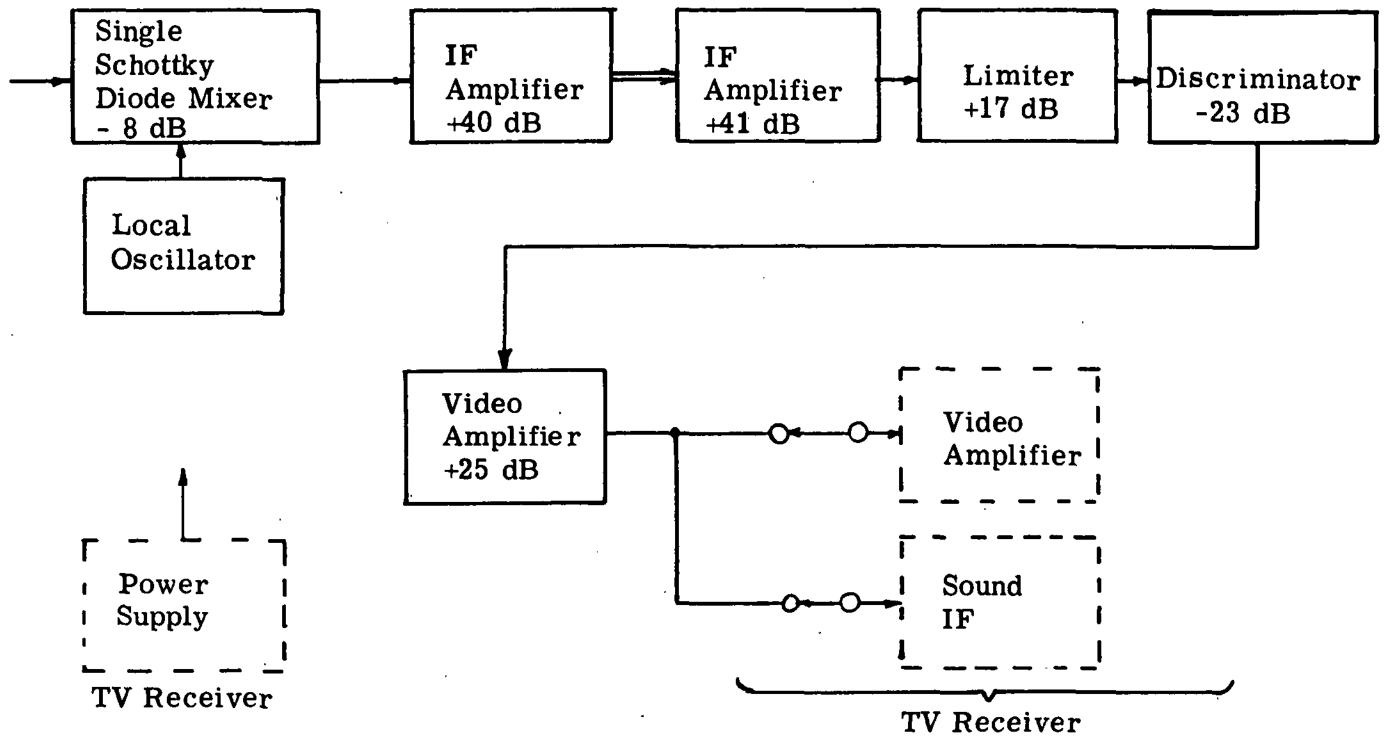


Figure 27. 12.0 GHz "Target-Cost" Converter Candidate System

It was assumed that significant cost reduction could be realized by

(1) utilizing a single diode Schottky mixer, due to the availability of excess local oscillator power from the Gunn oscillator.

(2) elimination of converter power supply and enclosure.

(3) elimination of the remodulator and oscillator circuitry of Figure 26.

(4) reduction of the "quality" of antenna unit gain, limiting, and discriminator linearity.

These proposed cost reductions would require the following modifications to the T. V. receiver:

(1) acquisition of dc power from the receiver power supply.

(2) insertion of a switch in the video and sound IF chain of the receiver to switch from satellite to VHF/UHF operation.

The power gain for each block is indicated in Figure 27. The output level and load impedance of the final video amplifier was chosen so that it is directly compatible with the General Electric K. E. chassis, and the -106 dBW signal level at the converter input is the specified input for "passable" performance.

The gains specified for each block in the converter were assumed to be readily achievable based on results obtained with the 12 GHz converter.

The use of a single Schottky diode instead of balanced mixer was a prime contender in converter cost reduction. Since a Gunn effect oscillator can deliver up to 20 mw power output, loose coupling of the L. O. signal to the mixer can be used without substantially affecting the signal. A design of a breadboard is shown in Figure 28, where a 10 dB directional coupler of $\frac{1}{2}$ wave length is used to achieve this objective. The IF short and the RF short are similar to that of the balanced structure used before. A microstrip design on 25 mil alumina substrate was used as a step toward better integration with the cavity. The overall assembly cost could be lower than the use of stripline structure.

Figure 29 is a block diagram of a second candidate system that was con sidered, whose principle of operation is quite different from the converters shown in Figures 26 and 27. The system incorporates F. M. feedback (FMFB) to reduce the bandwidth required of the IF amplifiers, limiters, and discriminators. Thus, these circuits can offer more gain per stage and the number of stages required to obtain the given video output is reduced.

The principle of operation of the system is quite simple. The center frequency of the local oscillator is offset from the center frequency of the RF carrier by the IF frequency. As the frequency of the RF shifts (due to modulation), the video output of the discriminator frequency modulates the local oscillator such that the frequency excursion of the L. O. approximately equals the frequency excursion of the R. F. carrier. The result is that for

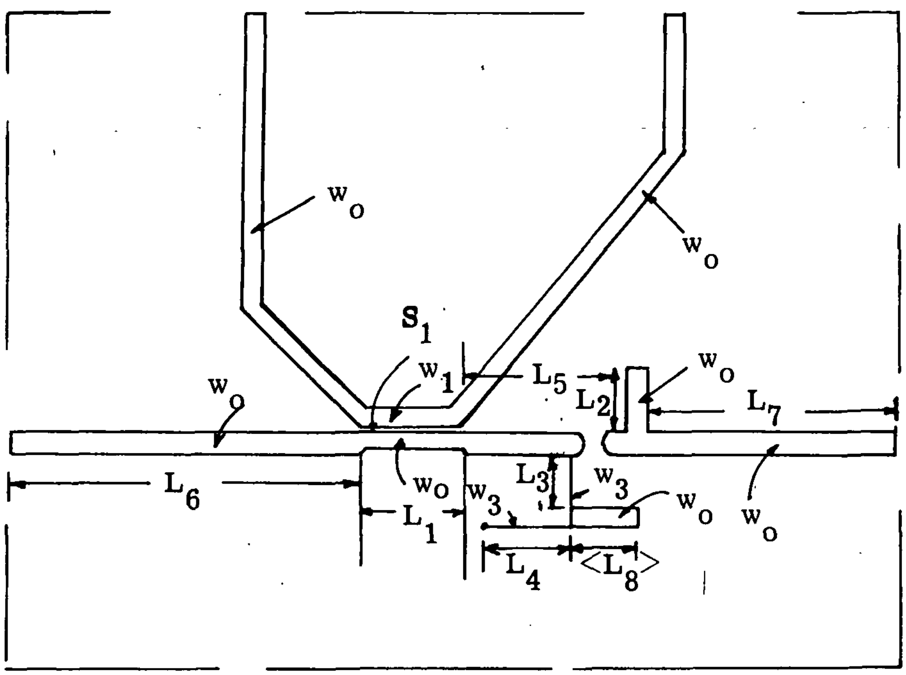


Figure 28. Single Diode Mixer Circuit for Target Cost Converter

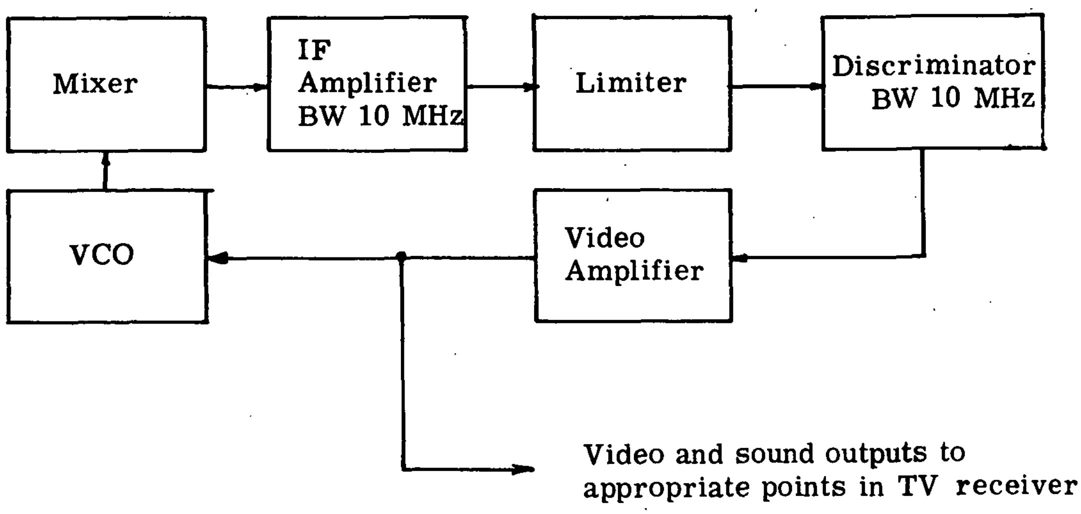


Figure 29. 12.0 GHz FMFB Converter Block Diagram

sufficiently high, open-loop gain, the IF frequency excursion is very small, and the IF signal can be considered to be narrow band FM. The bandwidth requirements for a narrow band FM signal is no greater than that required for a double sideband amplitude modulated signal. Thus, assuming that the highest modulated frequency is 4.5 MHz, an IF bandwidth of about 10 MHz is required by the FMFB system. This is considerably less than the 40 MHz bandwidth required by the IF amplifiers of the converters previously considered. Several difficulties are encountered in an FMFB system as just described. First, a wideband high gain feedback system must be carefully designed to insure stable operation. Secondly, the local oscillator must be capable of being linearly modulated at a high rate (4.5 MHz) and must be able to follow the full frequency shift of the RF carrier, which is about ±13 MHz.

It is conceivable that a Gunn diode could perform the VCO function required in the FMFB approach. This approach was discarded because the level of circuit development required was beyond the resources available for this task.

A third candidate was also considered. Some preliminary work was done to significantly reduce the cost of the microwave portion of the 12 GHz converter. It is conceivable to utilize the nonlinearities inherent in the Gunn diode local oscillator to perform the mixing function previously performed by the Schottky diode mixer. Some preliminary tests were done to determine the feasibility of this approach. Figure 30 is a simplified block diagram of the self oscillating mixer that was tested. The RF carrier was introduced into the cavity via a simple probe, the supply voltage was adjusted to set the oscillator 120 MHz below the carrier frequency, and the resulting IF was fed into a three stage IF amplifier previously designed as an IF preamplifier for the converter antenna unit. The IF amplifier had a center frequency of 120 MHz and a bandwidth of 40 MHz.

Measurements on the circuit of Figure 30 indicate that a signal-to-noise ratio of 1 was obtained with an RF input of -92 dBW. Since the converter must accommodate levels as low as -106 and offer a signal-to-noise ratio of the order of 10 dB, the noise figure of the system shown in Figure 30 must be improved by approximately 24 dB to satisfy the requirements of the present application. Although this is quite discouraging, it should be pointed out that these tests were intended to show feasibility and the circuit was not optimized.

Of the design alternatives considered, the first was selected for implementation. This approach was the only one that did not entail a level of circuit development beyond the resources available for the project.

Updated cost analyses were performed to determined the cost estimates for the original 12 GHz converter design and the target cost approach. The two cost breakdowns indicated that the major cost of the converters, the front end oscillator and mixer circuits, would be of the order of $16.00 in annual production quantities of a million units. Of this amount, the oscillator circuit (Gunn diode oscillator, cavity and voltage regulator) represented over 60% of the cost and it would be required in either approach. A savings of about $1.50 per unit, or less than 10% of the front end cost, could be obtained using a single diode mixer and a savings of another $1.00 could be realized by reducing the IF amplifier-limiter-discriminator circuit performance. The major cost savings would result from the elimination of the AM remodulator function, some power supply circuits, and the indoor unit enclosure. Some of these savings would be cancelled by the necessary modifications to the TV receiver. The net cost savings in this area would amount to about $4.00 per unit.

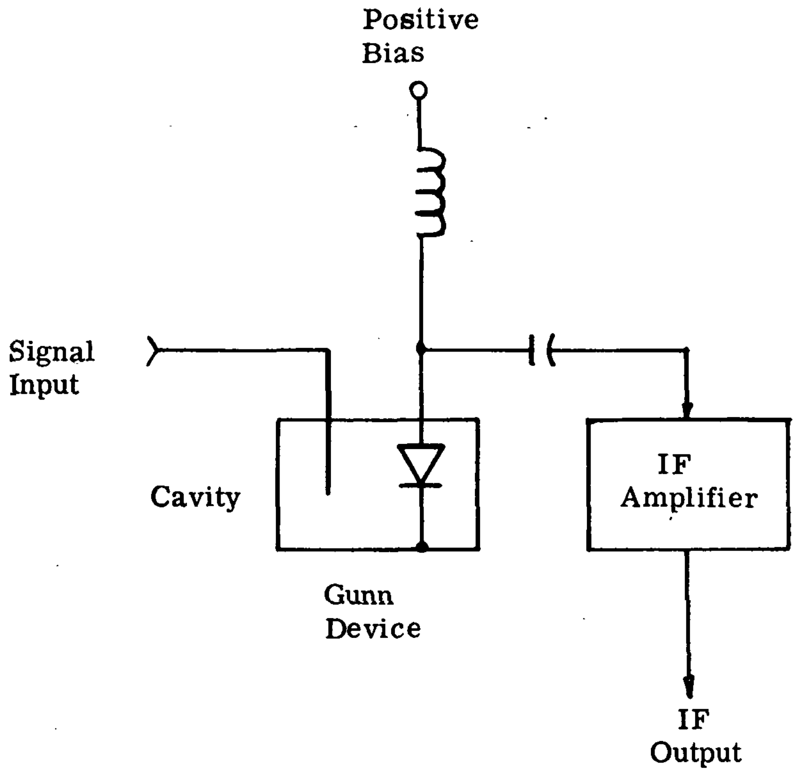


Figure 30. Test Configuration of Self-oscillating X-band Mixer

In total the cost of a 12 GHz FM converter could be reduced, at most, by 25%, compared to the target cost bogey of 50%. It was decided at this point to discontinue the target cost program since the expected cost savings would not justify the reduction in performance or increase in input signal level.

5.0 COST SENSITIVITY ANALYSIS

The scope of the contract was increased by adding TASK XIII entitled, "Cost Sensitivity Analysis". The objective of this task was to determine the influence of two performance parameters on the cost of the three converter types being developed. The two parameters, output signal-to-noise ratio and input signal level, were taken as variables independently. The output signal-to-noise ratio of the converters was varied, in one decibel steps, ten decibels below and above the original 35 dB specification. The input signal level to the converters was varied from eight decibels below to fifteen above the originally prescribed levels for the Type A and Type B converters. For the Type D converter the input signal level range was varied fron nine decibels below to fifteen above the original standard. The modulation index in the FM systems was allowed to vary to a maximum of three.

The output of the task were block diagrams and factory cost estimates for each parameter increment and converter type. The factory cost estimates were derived for four annual production levels, 10^3, 10^4, 10^5, and 10^6.

Code numbers were assigned to each converter type and each set of design parameters. For example, A-1 through A-20 designate the Type A converters whose output signal to noise ratios were specified from 25 dB to 45 dB with the input signal level, P_S, fixed at -92 dBW. The code number, A-0, was assigned to the converter with the original set of performance specifications. Similarly, A-21 through A-43 are the code numbers for converters with a fixed output signal-to-noise ratio of 35 dB while the antenna output power level was incremented from -100 dBW to -77 dBW. The A-0 converter, of course, appears again in that sequence. An equivalent notation is used for the Type B and Type D converters.

The combinations of parameters and converter types imply a total of 131 systems to be considered for conceptual design and cost analysis. The practical design options available do not represent a continuous succession of performance degradation or improvement. Step discontinuities result from the elimination or inclusion of functional blocks; thus several systems may share a common design.

The starting point in the cost sensitivity analyses was the updated cost analyses of the base designs for each converter type. The factory cost updates reflected design alterations and component price changes from the original cost estimates prepared in Phase I of the contract. Tables V, VI, and VII contain the detailed base cost analyses for the type A, B, and D converters.

The work required in this task was the matching of designs to performance criteria. This work consisted essentially of finding minimum cost designs that provide adequate noise performance. The specified parameters, input signal level and output signal-to-noise ratio, determined what the adequate noise

TABLE V

2.25 GHz AM SYSTEM, P_S = -92 dBW

S/N = 35 dB

DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
Local Oscillator				
1 HP 0180 Step Recovery Diode	2.00	1.80	1.50	.95
2 each 2N4996	.58	.42	.30	.10
2 each 2N3866	2.00	1.40	1.10	.98
1-72 MHz Crystal	2.00	2.00	2.00	2.00
R, L, & C's	4.19	3.72	3.23	3.23
4 In^2 G10 PCB	.12	.07	.06	.06
9 In^2 PPO PCB	1.20	.70	.63	.63
	12.09	10.11	8.82	7.95
Mixer				
2 each HP2811	1.20	1.06	.88	.62
Connector	1.72	1.21	.86	.74
16 In^2 PPO PCB	2.00	1.25	1.14	1.14
	4.92	3.52	2.88	2.50
I. F. Amplifier				
1 2N4996	.29	.21	.15	.10
R, L, C	.34	.34	.34	.34
4 In^2, G10 PCB	.12	.07	.06	.06
	.75	.62	.55	.50
Misc. Parts				
Enclosure	1.20	.40	.40	.40
Hardware & Misc.	1.50	1.25	1.00	1.00
Freight & Spoilage	.82	.51	.42	.37
	3.52	2.16	1.82	1.77
Labor, Assembly, & Test	5.15	2.05	1.55	1.05
Antenna Unit Total Materials & Labor	26.43	18.46	15.62	13.77

TABLE V (Cont.)

Indoor Unit

Power Supply	10^3	10^4	10^5	10^6
Power Transformer	.40	.40	.40	.40
2 Rectifier Diodes	.32	.32	.32	.32
3 Transistors	1.05	.84	.54	.36
1 Regulator Transistor	.50	.50	.50	.50
R, L, & C's	.55	.55	.55	.55
	2.82	2.61	2.31	2.13
Mis. Parts				
9 In^2 PCB	.12	.12	.12	.12
1 Antenna Switch	.15	.15	.15	.15
1 Screw Terminal Strip	.15	.15	.15	.15
Enclosure	.30	.30	.30	.30
Fuse	.08	.08	.08	.08
Fuse Clip	.10	.10	.10	.10
Pilot Lamp	.04	.04	.04	.04
Pilot Lamp Socket	.12	.12	.12	.12
Line Cord	.16	.16	.16	.16
30' Twin Lead	.60	.60	.60	.60
Hardware & Misc.	.40	.25	.25	.25
Freight & Spoilage	.29	.18	.15	.13
	2.51	2.24	2.22	2.20
Labor, Assembly & Test	.55	.25	.25	.25
Indoor Unit Total Materials & Labor	5.88	5.10	4.78	4.58
Total, 2.25 GHz Conv., 1970 Prices	32.31	23.56	20.40	18.35

TABLE VI

2.25 GHz FM System
P_S = -107 dBW
(S/N) = 35 dB
M. I. = 2

DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit

Local Oscillator	10^3	10^4	10^5	10^6
1 L.O. Transistor	9.50	4.00	3.75	1.75
R, L, & C's	.31	.31	.31	.31
6.8 In^2 PCB	.21	.21	.21	.21
	10.02	4.52	4.27	2.27
Mixer				
2 each HP2811 Diodes	1.20	1.06	.88	.62
1 Connector	1.72	1.21	.86	.74
12 In^2 PPO Board	1.50	.94	.86	.86
	4.42	3.21	2.60	2.22
I. F. Pre-Amplifier				
3 each 2N4996	.63	.45	.30	.30
R, L, & C's	1.10	1.06	.97	.97
8.6 In^2 PCB	.26	.15	.13	.13
	1.99	1.66	1.40	1.40
Misc. Parts				
Enclosure	1.20	.40	.40	.40
Hardware & Misc.	1.50	1.25	1.00	1.00
Freight & Spoilage	.48	.30	.25	.22
	3.18	1.95	1.65	1.62
Labor, Assembly & Test	3.50	1.75	1.17	1.10
Antenna Unit Total Materials & Labor	23.11	13.09	11.09	8.61

TABLE VI (Cont.)

Indoor Unit	10^3	10^4	10^5	10^6
I. F. Amplifier, Limiter & Disc.				
6 each 2N4996	1.26	.90	.60	.60
2 each HP2811	1.20	1.06	.88	.62
R, L, & C	2.31	2.24	2.03	2.03
22.8 In^2 G10 PCB	.68	.39	.35	.35
	5.45	4.59	3.86	3.60
Video Amplifier-Modulator				
6 each 2N4996	1.26	.90	.60	.60
1 Clamp Diode	.14	.11	.11	.11
R, L, & C's	1.26	1.18	1.14	1.14
11 In^2 G10 PCB	.33	.19	.17	.17
	2.99	2.38	2.02	2.02
Power Supply				
4 each A14A Diodes	.64	.64	.64	.64
4 each Transistors	.42	.38	.36	.36
Power Transformer & R, L, C's	1.76	1.76	1.76	1.76
12.2 In^2 PCB	.36	.21	.19	.19
	3.18	2.97	2.95	2.95
Misc. Parts				
Antenna Switch	.15	.15	.15	.15
Screw Terminal Strip	.15	.15	.15	.15
Fuse	.08	.08	.08	.08
Fuse Clip	.10	.10	.10	.10
Pilot Lamp	.04	.04	.04	.04
Pilot Lamp Socket	.12	.12	.12	.12
Enclosure	.50	.50	.50	.50
Line Cord	.16	.16	.16	.16
30' Twin Lead	.60	.60	.60	.60
Hardware & Misc.	.90	.60	.45	.38
Freight & Spoilage	.72	.45	.37	.33
	3.52	2.95	2.72	2.61
Labor, Assembly & Test	4.12	2.84	2.05	2.06
Indoor Unit, Total Materials & Labor	19.26	15.73	13.61	13.24
Total Cost 2.25 GHz FM System P_S = -107 dBW	42.34	28.80	24.68	21.83

S/N = 35 dB, M.I. = 2

TABLE VII

12 GHz FM System

P_S = -105 dBW

S/N = 35 dB

M.I. = 3

DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
Local Oscillator				
1 Gunn Diode	33.00	22.00	8.00	2.75
1 Cavity	5.95	3.22	2.19	2.01
1 Series Regulator Transistor	.50	.50	.50	.50
2 Transistors	.21	.18	.18	.18
R, L, C	.17	.17	.17	.17
	39.83	26.07	11.04	5.61
Mixer				
2 HP 2740	11.00	9.50	9.00	7.50
1 Connnector	1.72	1.21	.86	.74
8 In^2 PPO PCB	1.00	.63	.57	.57
	13.72	11.34	10.43	8.81
I. F. Pre Amp				
3 2N4996	.63	.45	.30	.30
R, L, C	1.10	1.06	.97	.97
8.6 In^2 PCB (G10)	.26	.15	.13	.13
	1.99	1.66	1.40	1.40
Misc. Parts				
Enclosure	1.20	.40	.40	.40
Hardware & Misc.	1.50	1.25	1.00	1.00
Freight & Spillage	1.80	.74	.60	.53
	4.50	2.39	2.00	1.93
Labor, Assembly, & Test	4.12	2.06	1.38	1.29
Antenna Unit Total Materials & Labor	64.16	43.52	26.25	19.04

TABLE VII (Cont.)

Indoor Unit	10^3	10^4	10^5	10^6
I. F. Amplifier, Limiter, Disc.				
6 each 2N4996	1.26	.90	.60	.60
2 each HP2811	1.20	1.06	.88	.62
R, L. C	2.31	2.24	2.03	2.03
22.8 In^2 PCB	.68	.39	.35	.35
	5.45	4.59	3.86	3.60
Video Amplifier-Modulator				
6 each 2N4996	1.26	.90	.60	.60
1 Clamp Diode (1N914)	.14	.11	.11	.11
R, L, C	1.26	1.18	1.14	1.14
11 In^2 PCB (G10)	.33	.19	.17	.17
	2.99	2.38	2.02	2.02
Power Supply				
4 each A14A Diodes	.64	.64	.64	.64
4 each Transistors	.42	.36	.36	.36
Power Transformer & R, L, C	2.27	2.27	2.27	2.27
12.2 In^2 PCB	.36	.21	.19	.19
	3.69	3.48	3.46	3.46
Misc. Parts				
Antenna Switch	.15	.15	.15	.15
Screw Terminal Strip	.15	.15	.15	.15
Fuse	.08	.08	.08	.08
Fuse Clip	.10	.10	.10	.10
Pilot Lamp	.04	.04	.04	.04
Pilot Lamp Socket	.12	.12	.12	.12
Line Cord	.16	.16	.16	.16
30' Twin Lead	.60	.60	.60	.60
Enclosure	.50	.50	.50	.50
Hardware & Misc.	.90	.60	.45	.38
Freight & Spoilage	.71	.44	.37	.32
	3.51	2.94	2.72	2.60
Labor, Assembly, Test	4.12	2.84	2.06	2.06
Indoor Unit Total Materials & Labor	19.76	16.23	14.12	13.74
Total Cost, 12 GHz FM System, P_A = -105 dBW, S/N = 35 dB, M.I. = 3	83.92	59.75	40.37	32.78

performance was for each case. Other performance parameters had to be simultaneously fulfilled. The converters had to meet interface needs such as impedance levels and signal format and provide adequate signal bandwidth and linearity. Most design changes occurred in the front-end circuits where the converter noise figure was established. The relationship for cascaded stages applied in determining the converter or system noise figure, i.e.,

$$F = F_1 + \frac{F_2-1}{G_1} + \frac{F_3-1}{G_1 \times G_2} + \ldots + \frac{F_n-1}{G_1 \times G_2 \times \ldots G_{n-1}}$$

where;

$F_{1,2,3,\ldots n}$ noise factors (power ratios) for individual stages

$G_{1,2,3,\ldots n-1}$ power gain ratios for individual stages where G_i may be less than unity.

Where possible the noise factor and gain ratios represent measured performance of blocks in the base systems.

The Type A converter systems are the simplest and consist of frequency translators between the microwave input signal and a standard VHF television channel frequency. The FM systems, Type B and Type D, require a frequency translation first to obtain low cost IF gain prior to the FM demodulation. The FM demodulator must operate at or above a minimum carrier-to-noise ratio to realize the FM improvement in signal to noise, I_{FM}. An additional variable, the modulation index of the FM signal, affects the amount of FM improvement that can be obtained, as well as the signal bandwidth required ahead of the FM demodulation. Regardless of the complexity added and the FM demodulator constraints, adequate noise performance at minimum cost is still the design goal for the FM systems.

5.1 TYPE A CONVERTER SYSTEMS (2.25 GHz AM)

The approach taken in this portion of the task was to establish the noise performance required by the base design for the Type A converter and to determine the limits of performance imposed by the variable parameter. The base design, A-0, was developed to provide an output signal-to-noise ratio of 35 dB with an antenna output power level of -92 dBW at sync tip. A correction factor is required when discussing output signal-to-noise ratio. The video output signal-to-noise ratio, S_o/N_o, is defined by the contract as the power ratio of the "peak-to-peak" video signal to the rms noise amplitude. The NTSC modulation format established the "peak-to-peak" video, or the white-to-blanking excursion, as 62.5% of the carrier at sync tip level. Therefore, the detected signal level available to contribute to the output signal-to-noise is only 0.625/0.707 times the rms carrier amplitude at sync tip. This represents a loss of 1.06 dB. The allowable noise degradation in the A-0 converter system is the difference between the input signal-to-noise ratio and the specified output signal-to-noise ratio:

$$\text{Noise Margin} = \frac{P_s^*}{P_n} - \frac{S_o}{N_o}$$

$P_s^* = P_s - 1.06 \text{ dB}$

P_n = Standard noise power in 4.5 MHz bandwidth

For P_s equal to -92 dBW,

$$\text{Noise Margin} = \frac{-92.0 \text{ dBW} - 1.06 \text{ dB}}{-147.44 \text{ dBW}} - 35 \text{ dB}$$

$$\text{Noise Margin} = 9.38 \text{ dB}$$

The noise margin corresponds to the maximum allowable noise figure of the A-0 converter system. The limits on allowable Type A converter system noise figure run from -0.62 dB to +24.38 dB. The lower limit is impossible because the output signal-to-noise specification exceeds the input signal-to-noise ratio.

The next step in the procedure is to generate minimum cost approaches to match these noise figure goals.

Tables VIII and IX summarize the specifications as well as the design and performance parameters for the Type A converter systems. As the specified output signal-to-noise was lowered from the A-0 requirement as in A-10, A-9, A-8, etc., the mixer and/or the IF amplifier performance degraded. Figure 31 is the A-0 block diagram with performance parameters. It was assumed that the television receiver performance specifications would be unchanged.

Modification #1: Substitution of Single Diode Mixer for the Balanced Mixer

The objective of this modification is to eliminate the cost of one mixer diode and a portion of the printed circuit board material required in the mixer. It is assumed that local oscillator injection can be performed without any significant increase in local oscillator power, and that signal port to local oscillator port isolation can be maintained at the present level. It is further assumed that the single diode mixer conversion gain will be the same as for the balanced mixer and that the mixer noise figure will increase by three decibels. The block diagram for Mod. #1 is shown in Figure 32 and the cost analysis is presented in Table X. The total converter system noise figure for this modification is 11.1 dB.

Modification #2: Elimination of the IF Amplifier

The elimination of the IF amplifier results in an increase of the system noise figure from two factors. First, the twinlead loss is added to the mixer loss because it appears in the system prior to the first gain stage. Second, the system noise figure is degraded because the noise figure of the television

TABLE VIII

FACTORY COST ESTIMATES - TYPE A CONVERTERS

2.25 GHz AM

Code No.	Specified S_o/N_o dB	F_T dB	BW MHz	P_N dBW	P_S/P_N* dB	S_o/N_o	Mod. No.	Annual Production Level 10^3	10^4	10^5	10^6
A-1	25	15.8	4.5	-137.4	44.3	28.5	3	29.49	21.67	18.53	16.30
2	26	15.8	4.5	-137.4	44.3	28.5	3	29.49	21.67	18.53	16.30
3	27	15.8	4.5	-137.4	44.3	28.5	3	29.49	21.67	18.53	16.30
4	28	15.8	4.5	-137.4	44.3	28.5	3	29.49	21.67	18.53	16.30
5	29	15.4	4.5	-137.4	44.3	28.9	2	30.71	22.59	19.59	17.68
6	30	11.1	4.5	-137.4	44.3	33.2	1	31.09	22.64	19.61	17.69
7	31	11.1	4.5	-137.4	44.3	33.2	1	31.09	22.64	19.61	17.69
8	32	11.1	4.5	-137.4	44.3	33.2	1	31.09	22.64	19.61	17.69
9	33	11.1	4.5	-137.4	44.3	33.2	1	31.09	22.64	19.61	17.69
10	34	9.4	4.5	-137.4	44.3	34.9		32.31	23.56	20.40	18.35
0	35	9.4	4.5	-137.4	44.3	34.9					
11	36	7.2	4.5	-137.4	44.3	37.1	4	51.83	33.71	29.31	25.15
12	37	7.2	4.5	-137.4	44.3	37.1	4	51.83	33.71	29.31	25.15
13	38	5.0	4.5	-137.4	44.3	39.3	4a	101.83	61.21	53.31	44.15
14	39	5.0	4.5	-137.4	44.3	39.3	4a	101.83	61.21	53.31	44.15
15	40	2.6	4.5	-137.4	44.3	41.7	5	264.17	201.89	125.44	82.42
16	41	2.6	4.5	-137.4	44.3	41.7	5	264.17	201.89	125.44	82.42
17	42	2.6	4.5	-137.4	44.3	41.7	5	264.17	201.89	125.44	82.42
18	43	(1.3)	4.5	-137.4	44.3	----		264.17	201.89	125.44	82.42
19	44	(0.3)	4.5	-137.4	44.3	----		Cryogenic > $7,500			
20	45	—	4.5	-137.4	44.3	----	-				

P_S/P_N* Corrected to account for difference between defined output S/N and rms S/N.

P_S = -92 dBW Signal power delivered to converter input from the antenna at sync tip with a matched load.

TABLE IX

FACTORY COST ESTIMATES - TYPE A CONVERTERS

2.25 GHz AM

Code No.	P_s dBW	F_T dB	BW MHz	P_N dBW	P_S/P_N* dB	S_o/N_o dB	Mod. No.	Annual Production Level (1970) 10^3	10^4	10^5	10^6
A-21	-100	(1.3)	4.5	-137.4	36.3	----	-	Cryogenic		7,500	
22	- 99	(2.3)	4.5	-137.4	37.3	----	-				
23	- 98	2.6	4.5	-137.4	38.3	35.7	5	264.17	201.89	125.44	82.42
24	- 97	2.6	4.5	-137.4	39.3	36.7	5	264.17	201.89	125.44	82.42
25	- 96	5.0	4.5	-137.4	40.3	35.3	4a	101.83	61.21	53.31	44.15
26	- 95	5.0	4.5	-137.4	41.3	36.3	4a	101.83	61.21	53.31	44.15
27	- 94	7.2	4.5	-137.4	42.3	35.1	4	51.83	33.71	29.31	25.15
28	- 93	7.2	4.5	-137.4	43.3	36.1	4	51.83	33.71	29.31	25.15
0	- 92	9.4	4.5	-137.4	44.3	34.9		32.31	23.56	20.40	18.35
29	- 91	9.4	4.5	-137.4	45.3	35.9		32.31	23.56	20.40	18.35
30	- 90	11.1	4.5	-137.4	46.3	35.2	1	31.09	22.64	19.61	17.69
31	- 89	11.1	4.5	-137.4	47.3	36.2	1	31.09	22.64	19.61	17.69
32	- 88	11.1	4.5	-137.4	48.3	37.2	1	31.09	22.64	19.61	17.69
33	- 87	11.1	4.5	-137.4	49.3	38.2	1	31.09	22.64	19.61	17.69
34	- 86	15.4	4.5	-137.4	50.3	34.9	2	30.71	22.59	19.59	17.68
35	- 85	15.8	4.5	-137.4	51.3	35.5	3	29.49	21.67	18.53	16.30
36	- 84	15.8	4.5	-137.4	52.3	36.5	3	29.49	21.67	18.53	16.30
37	- 83	15.8	4.5	-137.4	53.3	37.5	3	29.49	21.67	18.53	16.30
38	- 82	15.8	4.5	-137.4	54.3	38.5	3	29.49	21.67	18.53	16.30
39	- 81	15.8	4.5	-137.4	55.3	39.5	3	29.49	21.67	18.53	16.30
40	- 80	15.8	4.5	-137.4	56.3	40.5	3	29.49	21.67	18.53	16.30
41	- 79	15.8	4.5	-137.4	57.3	41.5	3	29.49	21.67	18.53	16.30
42	- 78	15.8	4.5	-137.4	58.3	42.5	3	29.49	21.67	18.53	16.30
43	- 77	15.8	4.5	-137.4	59.3	43.5	3	29.49	21.67	18.53	16.30

TABLE X

MODIFICATION #1 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
Local Oscillator				
Same as present	12.09	10.11	8.82	7.95
Mixer				
1 each HP 2811	.60	.53	.44	.31
Connector	1.72	1.21	.86	.74
11 In^2 PPO	1.38	.86	.79	.79
(~2/3 Balanced Mixer area)				
	3.70	2.60	2.09	1.84
IF, Misc.				
Labor, Assembly, & Test				
Same as Present	9.42	4.83	3.92	3.32
Antenna Unit, Total Costs, Materials & Labor	25.21	17.54	14.83	13.11
Indoor Unit Total (Same as Present)	5.88	5.10	4.78	4.58
Total Costs, Mod #1	31.09	22.64	19.61	17.69

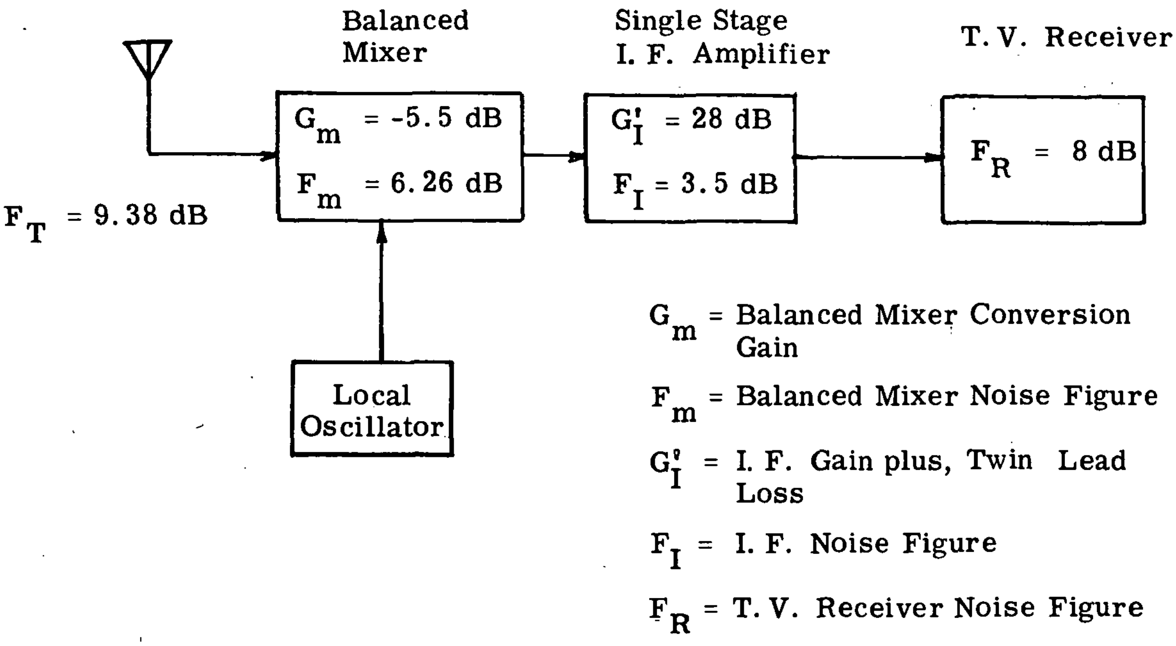


Figure 31. A-zero Simplified Block Diagram

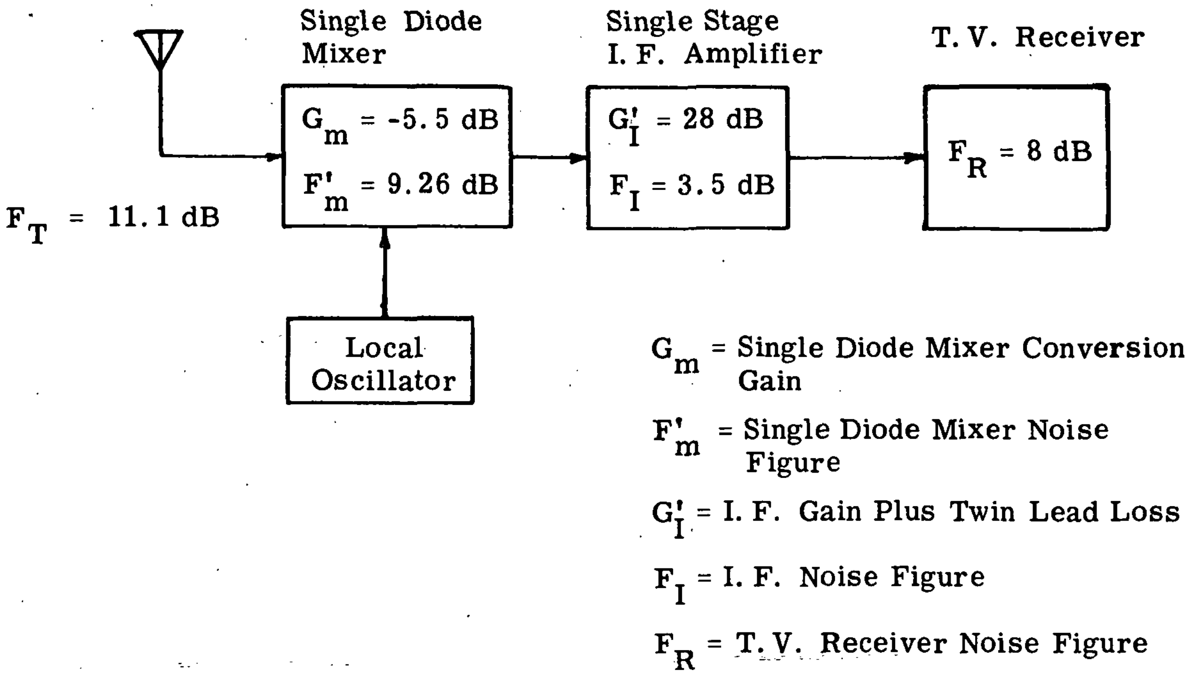


Figure 32 . Modification #1 Simplified Block Diagram

receiver is greater than the noise figure of the IF amplifier that is removed. The block diagram for Mod. #2 is shown in Figure 33 and the cost analysis in Table XI. The total system noise figure for this modification is 15.35 dB.

Modification #3: Single Diode Mixer with no IF Amplifier

This modification represents the minimum configuration required for the basic converter function of frequency translation. The incremental cost savings are minimal, which indicates that increasing the signal power is not a good design trade-off. The block diagram for Mod #3 is shown in Figure 34, the cost analysis in Table XII. The system noise figure is 15.85 dB.

Modification #4: Addition of a Transistor RF Amplifier

Modification #4 provides for a reduction of the system noise figure for those with either a lower input signal level or a higher output signal-to-noise requirement. The RF amplifier represents a volatile area regarding device cost and performance. A great deal of development has occured in microwave transistors in recent years and it is expected that further improvements will occur in the near future. The performance assigned to the RF amplifier at the time of the analysis reflects the best trade-off of performance and cost. The block diagram for this modification is shown in Figure 35. The cost analysis for the modification is presented in Table XIII. The system noise figure using this modification is 7.16 dB. It is felt that a further improvement in the system noise figure can be realized with the microwave transistors that are available today. A system noise figure of 5.0 dB is possible with an additional incremental cost of about $50. This option will be called Modification #4a and has the same block diagram as shown in Figure 35.

Modification #5: Addition of a Parametric RF Amplifier

Figure 36 shows the block diagram for the parametric amplifier used in modification #5. The 2.25 GHz input signal is up-converted to the 12 GHz. region to achieve a gain of 15 dB with a 1.6 dB noise figure. The X-band output is down-converted to the IF or output frequency as before. Special care is required in designing the pump and local oscillator circuits to eliminate any spurious FM that would interfere with the aural subcarrier signal in the transmitted format. This would be achieved by using double cavity oscillators with an estimated incremental cost of 50% over the X-band sources used in the Type D converters. The total system noise figure using the parametric preamplifier would be 2.6 dB. The cost analysis is contained in Table XIV.

Further Noise-Figure Improvement

There are five Type A converter specifications that cannot be met with the noise figure realized with an uncooled parametric preamplifier. Systems operating to these specifications will not have a place in a mass-produced product. A rough estimate of the cost of a cryogenic system is in excess of $7,000. This does not include electronic circuit costs and antenna modifications. Detailed cost estimates for such systems were not made under the present program.

TABLE XI

MODIFICATION #2 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
Local Oscillator				
Same as Present	12.09	10.11	8.82	7.95
Balanced Mixer				
Same as Present	4.92	3.52	2.88	2.50
Misc. Parts				
Same as Present	3.52	2.16	1.82	1.77
Labor, Assembly & Test	4.30	1.70	1.29	.88
Total Antenna Unit Costs	24.83	17.49	14.81	13.10
Total Indoor Unit Costs				
(Same as Present)	5.88	5.10	4.78	4.58
Total Costs, Mod #2	30.71	22.59	19.59	17.68

TABLE XII

MODIFICATION #3 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
Local Oscillator				
(Same as Present)	12.09	10.11	8.82	7.95
Single Diode Mixer				
(Same as MOD #1)	3.70	2.60	2.09	1.84
Misc. Parts				
(Same as Present)	3.52	2.16	1.55	1.05
Labor, Assembly & Test				
(Same as MOD #2)	4.30	1.70	1.29	.88
Total Antenna Unit Costs	23.61	16.57	13.75	11.72
Indoor Unit Costs				
(Same as Present)	5.88	5.10	4.78	4.58
Total Costs, Mod #3	29.49	21.67	18.53	16.30

TABLE XIII

MODIFICATION #4 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
R. F. Amplifier				
Transistor (T.I. MS173)	16.00	8.00	7.00	5.00
16 In^2 PPO PCB	2.00	1.25	1.14	1.14
R, L, & C's	.34	.34	.34	.34
Misc, Freight & Spoilage	.33	.21	.17	.15
Labor	.85	.35	.26	.17
Total RF Amplifier Costs	19.52	10.15	8.91	6.80
Mixer, Local Oscillator, IF Amplifier, etc.				
(Same as present antenna unit costs)	26.43	18.46	15.62	13.77
Total Antenna Unit Costs	45.95	28.61	24.53	20.57
Indoor Unit Costs				
(Same as Present)	5.88	5.10	4.78	4.58
Total Costs, Mod #4	51.83	33.71	29.31	25.15

TABLE XIV

MODIFICATION #5 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
X-Band Pump				
(Same as XFM-0 L.O.)	39.83	26.07	11.04	5.61
Varactor Diode Amp				
Microwave Associates MA-4536	105.00	95.00	67.00	42.50
16 In^2 PPO	2.00	1.26	1.14	1.14
X-Band Mixer				
(Same as XFM-0 Mixer)	13.72	11.34	10.43	8.81
X-Band Local Oscillator				
(Same as XFM-0 L.O.)	39.83	26.07	11.04	5.61
IF Amplifier				
(Same As Present)	.75	.62	.55	.50
Misc. Parts				
Enclosure	1.20	.40	.40	.40
Hard. & Misc.	2.25	1.87	1.50	1.50
Fgt. & Spoilage				
Labor, Assembly & Test	8.24	4.12	2.76	2.58
Antenna Unit Total Cost*	217.22	169.50	108.12	70.55
Pump and L.O. Cavity Mod.	39.83	26.07	11.04	5.61
Antenna Unit Total Cost	257.05	195.57	119.16	76.16

TABLE XIV (continued)

Indoor Unit	10^3	10^4	10^5	10^6
Power Supply	4.06	3.83	3.81	3.81
Misc. Parts				
(Same as Present)	2.51	2.24	2.22	2.20
Labor Assembly & Test				
(Same as Present)	.55	.25	.25	.25
Total Indoor Unit Costs	7.12	6.32	6.28	6.26
Antenna Unit Total Cost	257.05	195.57	119.16	76.16
Total Costs, Mod #5	264.17	201.89	125.44	82.42

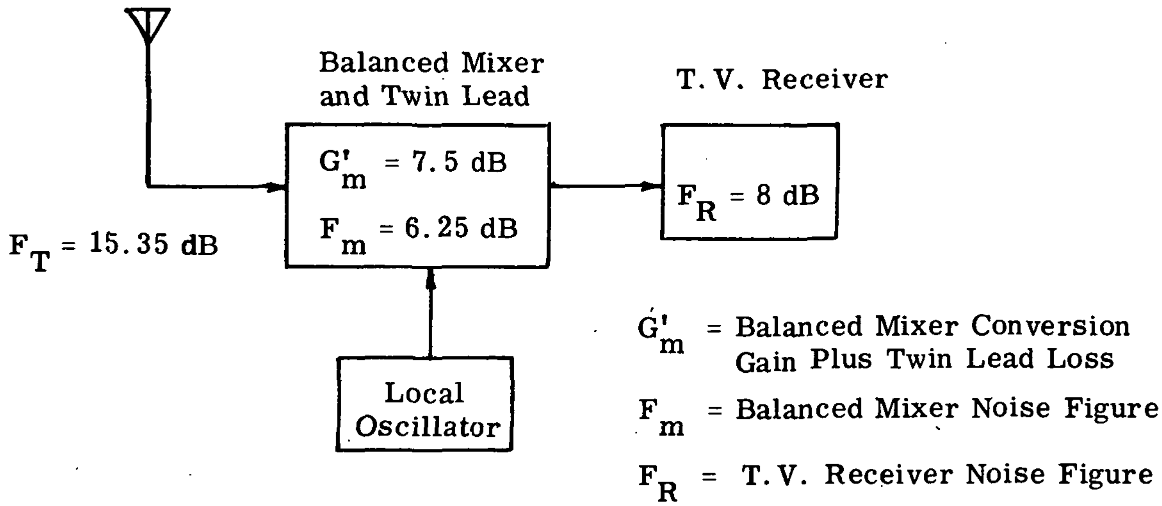


Figure 33 . Modification #2 Simplified Block Diagram

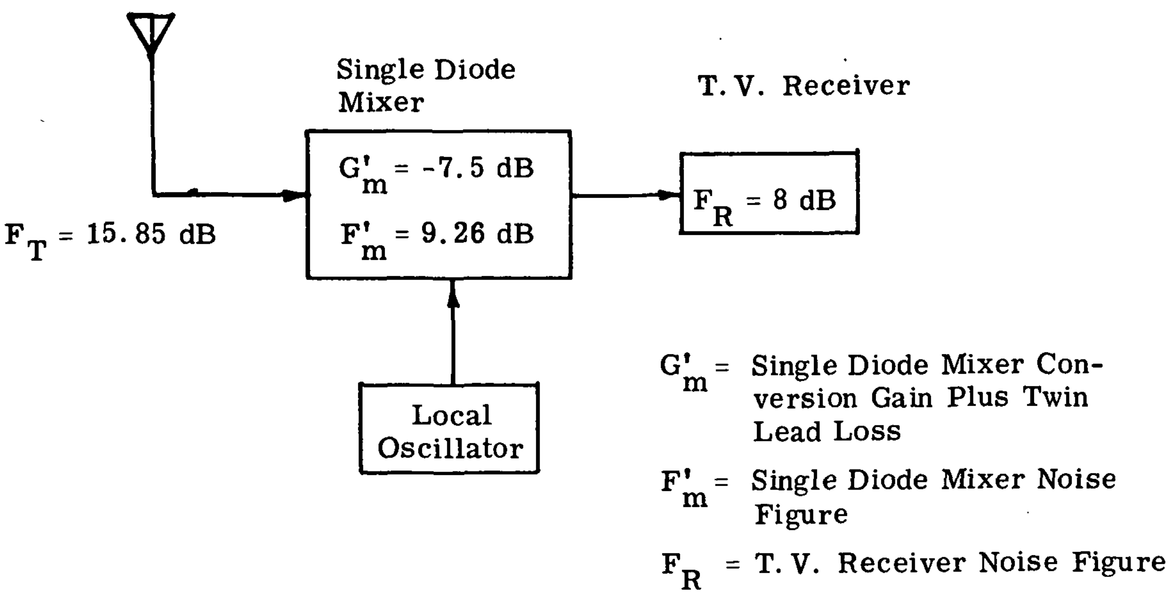


Figure 34 . Modification #3 Simplified Block Diagram

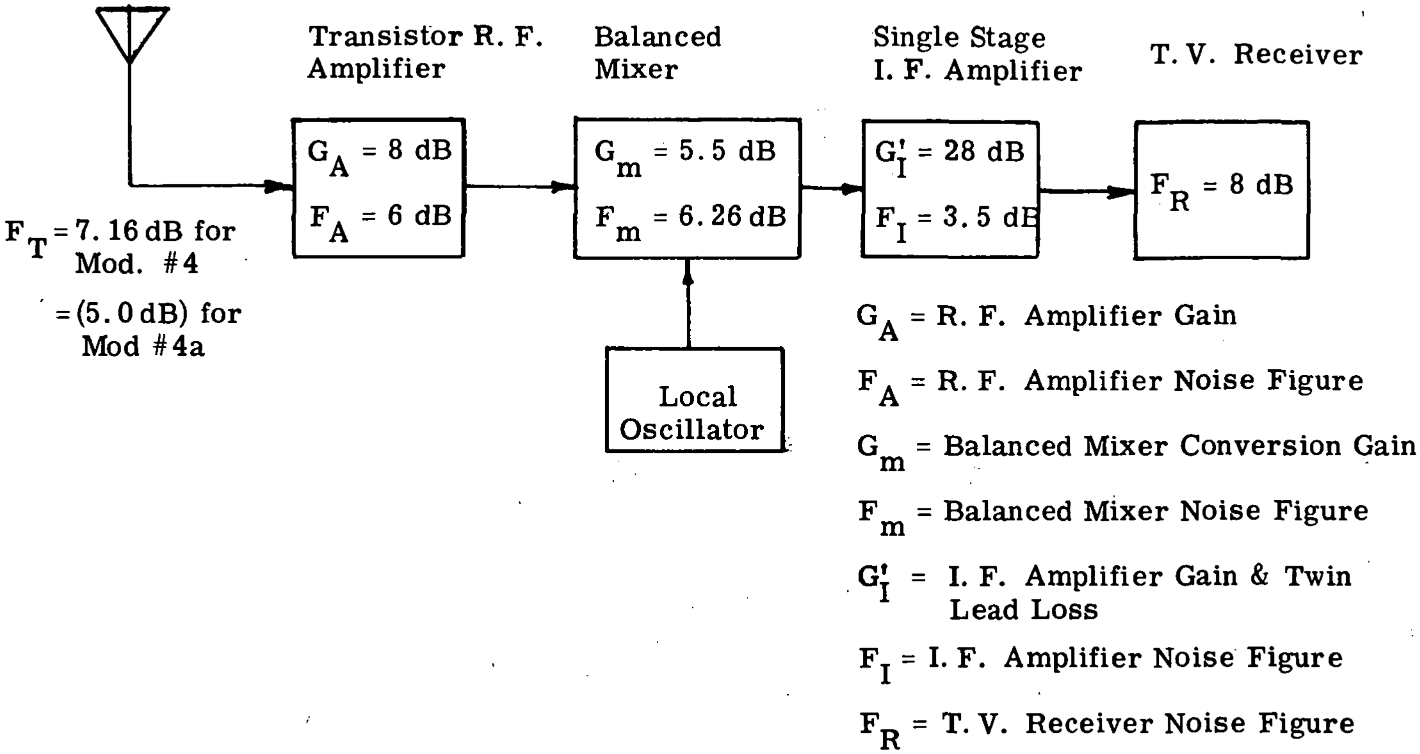


Figure 35. Modification #4 Simplified Block Diagram

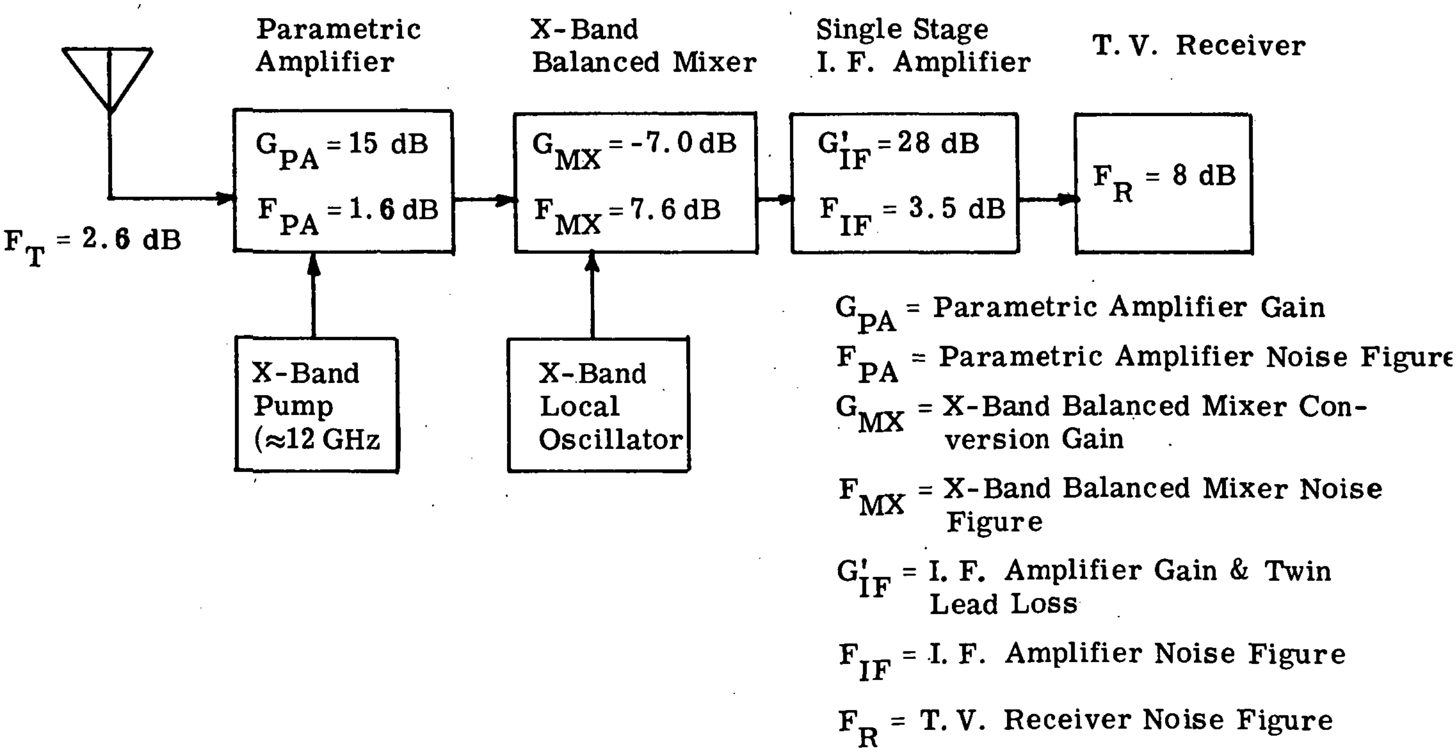


Figure 36. Modification #5 Simplified Block Diagram

5.2 TYPE B CONVERTER SYSTEMS (2.25 GHz FM)

The FM converter systems are distinct from the AM converters in that an additional parameter, the modulation index, can be used in the design to match system performance to system needs. The FM modulation index is defined as a ratio of the carrier deviation limits to the modulating frequency. The modulation index affects the RF bandwidth requirements and, thereby, affects the carrier-to-noise ratio at the discriminator input. A minimum carrier to noise ratio of 11.5 dB has been assumed to be adequate to insure the signal-to-noise improvement characteristic, I_{FM}, of an FM system. When this threshold carrier-to-noise ratio is maintained, the system output signal-to-noise ratio will increase with an increase in the modulation index. When complex modulating waveforms are used there are individual modulation indices for each spectral component of the waveform. To circumvent this complexity the modulation index resulting from a composite video signal and aural subcarrier signal is defined as the peak-to-peak carrier deviation caused by the white-to-sync tip video waveform divided by twice the highest modulating frequency. In this case the highest modulating frequency is the aural subcarrier or 4.5 MHz. When this definition is used the carrier deviation limits track the modulation index. Additional bandwidth is required over the peak-to-peak carrier deviation limits to accommodate the sideband energy resulting from the modulation process. Analysis has shown that an additional 4.5 MHz beyond each carrier deviation limit is sufficient to contain the significant sideband signals. The RF bandwidth allotted to the FM systems is a linear function of the modulation index.

$$\mathrm{BW} = 9\ \mathrm{MHz} \times \text{Modulation Index} + 2 \times 4.5\ \mathrm{MHz} + 3\ \mathrm{MHz}$$

The first term accommodates the peak-to-peak carrier deviation per the definition of the modulation index. The second term covers the important sideband energy, and the last term allows for local oscillator drift in the converter.

The output signal-to-noise ratio of the video signal is defined as the ratio of the peak-to-peak signal excursions between white and blanking video levels divided by the rms noise in the video bandwidth. The signal to noise calculated from the power relationships infers the rms value of the video signal whose peaks are at blanking and white levels. The specified output signal-to-noise is, therefore, reduced by 9.0 dB to be compatible with the power calculations. The rms output signal-to-noise relationship for the FM systems is as follows:

$$\left(S_o/N_o\right)_{RMS} = Ps(\mathrm{dBW}) - Pn\ (\mathrm{dBW}) - F_T(\mathrm{dB}) + I_{FM}(\mathrm{dB})$$

where:

- Ps = Signal power at the converter input
- Pn = Noise power in the converter IF
- F_T = Total converter system noise figure
- I_{FM} = FM noise improvement factor

I_{FM} and Pn are functions of the modulation index used

$$I_{FM} = 10 \log_{10} \left[6(M + 1)(0.765\ M)^2\right]$$

$$Pn = -137.44 \text{ dBW} + 10 \log_{10} \left[2(M+1) + 2/3\right]$$

where M = defined modulation index

The procedure used in the selection of the modulation index for the various FM converters is as follows. A plot of I_{FM} - Pn was generated as a function of the modulation index, M (see Figure 37). The magnitude of this quantity required for each set of converter specifications was found by rearranging the signal-to-noise relationship,

$$I_{FM} - Pn = (S_o/N_{o_{rms}} - Ps + F_T) \text{ dBW.}$$

The signal-to-noise and signal power values are given in the specifications and the noise figure value is obtained from the converter configurations considered. A check is made at each point to insure an adequate carrier-to-noise ratio at the discriminator input with the following relationships.

$$C/N_{rms} = (Ps - Pn - F_T) \text{ dB}$$

The value of modulation index required is found by entering the plot of I_{FM} - Pn versus M at the value determined from the relationship given above. This value of M determines the bandwidth and, therefore, the noise power, Pn. The prediscriminator carrier-to-noise ratio can then be checked to insure that the design is operating at or above the FM threshold.

Most of the converter designs will operate near the FM threshold to realize the minimum cost. The converters, whose input signal level or output signal-to-noise ratio are specified at the upper limits, will tend to operate with higher C/N, For those cases where a high output signal-to-noise ratio is specified, operation above the FM threshold is required because an upper limit of 3 was placed on the modulation index.

The Type B converter base design was developed to provide an output signal-to-noise ratio of 35 dB with an input signal level of -107 dBW using a modulation index of 2. Starting from this point, the improvement or degradation of the converter performance is adjusted to meet the parameter variations in the cost sensitivity analysis. The means used to improve or degrade performance are much the same as with the Type A converter. The lower system noise figures are obtained by adding low noise RF preamplifiers. There is little cost advantage to be obtained in the degradation of the system noise figure. The relative cost of the IF transistors, 2N4996, and those having the same $f_{\alpha b}$ with higher noise figure is not significant.

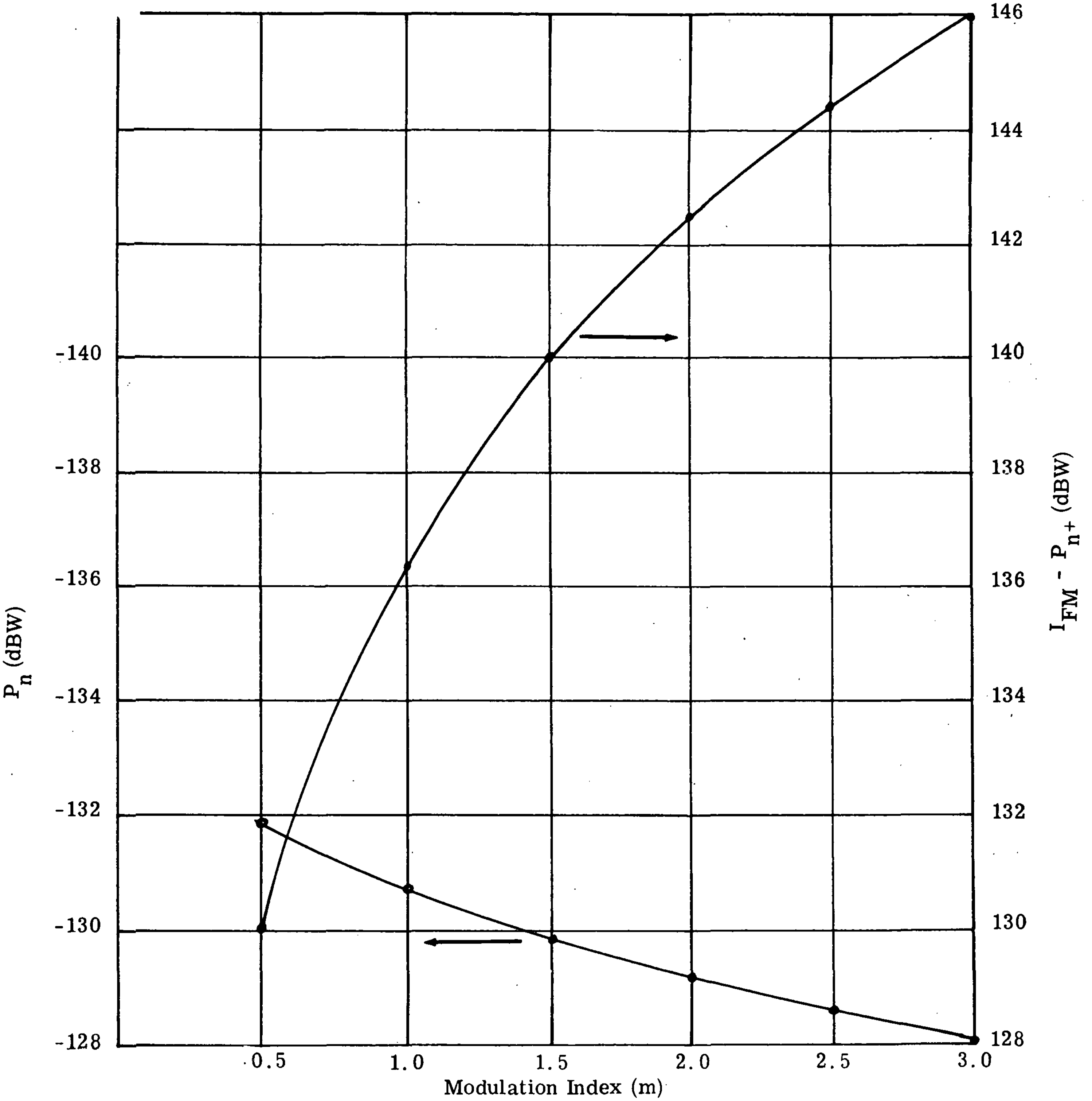


Figure 37. Net FM Improvement (I_{FM}-P_N) versus Modulation Index

The amount of gain required in the IF amplifier depends on the input signal level. Also, the IF bandwidth depends on the modulation index required. An analysis based on the gain - bandwidth capability of the IF transistor indicates that the IF amplifier in the FM converters requires from five to seven stages of gain depending on the specifications for the individual cases. The analysis assumes a maximum stable gain per stage of 17 dB and accounts for the fractional bandwidth loss in cascaded stagger-tuned stages from that obtained with one single-tuned stage. A plot of available IF gain versus IF bandwidth as a function of the number of IF stages is shown in Figure 38. The performance requirement for the IF amplifier slightly exceeds that of a seven stage IF for two FM converter configurations. It is felt that this gain shortage can be made up in the limiter circuit and that seven IF stages will be adequate for all cases.

Four modifications to the basic Type B converter design were considered as viable options, plus the modifications that alter the number of IF amplifier stages. Tables XV and XVI summarize the specifications as well as the design and performance parameters for the Type B converter systems.

Modification #6 <u>Substitution of a Single Diode Mixer for the Balanced Mixer</u>

This modification is essentially the same as Modification #1 for the Type A converter systems. The total system noise figure degrades to 11.2 dB in this case because of the lower gain per stage in the wideband FM intermediate frequency amplifier. Figure 39 shows the block diagram and performance parameters for the B-zero or base design for the Type B converters. Figure 40 shows the block diagram and parameters for the Modification #6 approach. The cost analysis for Modification #6 is contained in Table XVII. The system noise figure with Modification #6 is 11.2 dB.

Modification #7 <u>Elimination of IF Amplifier Stages</u>

This modification is an adjunct to the other modifications discussed for the FM converters. The block diagrams remain essentially unchanged as does the noise figure of the particular configuration. The net cost savings for the elimination of an IF stage is shown in Table XVIII. This savings is doubled when two stages are eliminated.

Modification #8 <u>Improvement of IF Amplifier Noise Figure</u>

A slight improvement in the IF noise figure permits a large increase in the output signal-to-noise of the Type B converter through the increase of the modulation index. The IF noise figure improvement maintains an adequate carrier to noise level at the discriminator as the bandwidth increases as a result of the modulation index increase.

The block diagram for this modification, Figure 41, is unchanged from the B-zero system except for the performance parameters. Substitution of a field effect transistor as the first IF amplifier provides the noise figure improvement that decreases the system noise figure to 8.4 dB. The cost analysis for Modification #8 is given in Table XIX.

TABLE XV

FACTORY COST ESTIMATES - TYPE B CONVERTERS

2.25 GHz FM

Code No.	S_o/N_o*	F_T dB	$I_{FM}-P_N$ dBw	M.I.	B.W. MHz	P_N dBW	C/N dB	MOD. No.	10^3	10^4	10^5	10^6
B-1	25	11.2	134.2	0.8	19	-131.1	12.9	6	41.62	28.19	24.17	21.45
2	26	11.2	135.2	0.9	20	-131.0	12.8	6	41.62	28.19	24.17	21.45
3	27	11.2	136.2	1.0	21	-130.7	12.5	6	41.62	28.19	24.17	21.45
4	28	11.2	137.2	1.1	22	-130.5	12.3	6	41.62	28.19	24.17	21.45
5	29	11.2	138.2	1.2	23	-130.3	12.1	6	41.62	28.19	24.17	21.45
6	30	11.2	139.2	1.4	25	-130.0	11.8	6	41.62	28.19	24.17	21.45
7	31	11.2	140.2	1.6	26	-129.8	11.6	6	41.62	28.19	24.17	21.45
8	32	9.5	139.5	1.4	25	-130.0	13.5		42.34	28.80	24.68	21.83
9	33	9.5	140.5	1.6	26	-129.8	13.3		42.34	28.80	24.68	21.83
10	34	9.5	141.5	1.8	28	-129.5	13.0		42.34	28.80	24.68	21.83
0	35	9.5	142.5	2.0	30	-129.2	12.7		42.34	28.80	24.68	21.83
11	36	9.5	143.5	2.2	32	-128.9	12.4		42.34	28.80	24.68	21.83
12	37	9.5	144.5	2.5	34	-128.6	12.1		42.34	28.80	24.68	21.83
13	38	9.5	145.5	2.8	37	-128.2	11.7		42.34	28.80	24.68	21.83
14	39	8.4	145.4	2.8	37	-128.2	12.8	8	43.34	29.40	25.03	22.18
15	40	7.2	145.2	2.8	37	-128.2	14.0	9	61.86	38.95	33.59	28.63
16	41	5.0	144.0	2.4	34	-128.7	16.7	9a	111.86	66.45	57.59	47.63
17	42	5.0	145.0	2.7	36	-128.4	16.4	9a	111.86	66.45	57.59	47.63
18	43	5.0	146.0	3.0	39	-128.0	16.0	9a	111.86	66.45	57.59	47.63
19	44	2.6	144.6	2.6	35	-128.5	18.9	10	238.57	186.91	133.42	86.21
20	45	2.6	145.6	2.9	38	-128.1	18.5	10	238.57	186.91	133.42	86.21

$S_o/N_o{}^* = \frac{\text{Peak to Peak Video}}{\text{RMS Noise}}$

P_S = - 107 dBW Signal power delivered to converter input from the antenna at sync tip with a matched load.

TABLE XVI

FACTORY COST ESTIMATE - TYPE B CONVERTER
2.25 GHz FM

Code No.	P_S dBW	F_T dB	$I_{FM}-P_N$ dBW	M.I.	BW MHz	P_N dBW	C/N dB	G_{IF} dB	N_{IF}	Mod. No.	Annual Production (1970) 10^3	10^4	10^5	10^6
B-21	-115	2.6	143.6	2.3	33	-128.8	11.2**	103	7	10	238.57	186.91	133.42	86.21
22	-114	2.6	142.6	2.1	31	-129.1	12.5	102	7	10	238.57	186.91	133.42	86.21
23	-113	2.6	141.6	1.8	28	-129.5	13.9	101	7	10	238.57	186.91	133.42	86.21
24	-112	5.0	143.0	2.2	32	-128.9	11.9	100	7	9a	111.86	66.45	57.59	47.63
25	-111	5.0	142.0	2.0	30	-129.2	13.2	99	7	9a	111.86	66.45	57.59	47.63
26	-110	7.2	143.2	2.2	32	-128.9	11.7	98	7	9	61.86	38.95	33.59	28.63
27	-109	7.2	142.2	2.0	30	-129.2	13.0	97	7	9	61.86	38.95	33.59	28.63
28	-108	8.4	142.4	2.0	30	-129.2	12.8	96	6	8	42.62	28.80	24.53	21.69
0	-107	9.5	142.5	2.0	30	-129.2	12.7	95	7		42.34	28.80	24.68	21.83
29	-106	11.2	143.2	2.2	32	-128.9	11.7	94	7	6	41.62	28.19	24.17	21.45
30	-105	11.2	142.2	2.0	30	-129.2	13.0	93	6	6	40.90	27.59	23.67	20.96
31	-104	11.2	141.2	1.8	28	-129.5	14.3	92	6	6	40.90	27.59	23.67	20.96
32	-103	11.2	140.2	1.6	26	-129.8	15.6	91	6	6	40.90	27.59	23.67	20.96
33	-102	11.2	139.2	1.4	25	-130.0	16.8	90	6	6	40.90	27.59	23.67	20.96
34	-101	11.2	138.2	1.2	23	-130.3	18.1	89	6	6	40.90	27.59	23.67	20.96
35	-100	11.2	137.2	1.1	22	-130.5	19.3	88	5	6	40.18	26.99	23.17	20.47
36	- 99	11.2	136.2	1.0	21	-130.7	20.5	87	5	6	40.18	26.99	23.17	20.47
37	- 98	11.2	135.2	0.9	20	-131.0	21.8	86	5	6	40.18	26.99	23.17	20.47
38	- 97	11.2	134.2	0.8	19	-131.1	22.9	85	5	6	40.18	26.99	23.17	20.47
39	- 96	11.2	133.2	0.7	18	-131.4	24.2	84	5	6	40.18	26.99	23.17	20.47
40	- 95	11.2	132.2	0.7	18	-131.4	25.2	83	5	6	40.18	26.99	23.17	20.47
41	- 94	11.2	131.2	0.6	17	-131.6	26.4	82	5	6	40.18	26.99	23.17	20.47
42	- 93	11.2	130.2	0.6	17	-131.6	27.4	81	5	6	40.18	26.99	23.17	20.47
43	- 92	11.2	129.2	0.5	16	-131.8	28.6	80	5	6	40.18	26.99	23.17	20.47

* S_o/N_o = 35 dB

**C/N is low but system cost does not warrant further NF reduction

G_{IF} = IF amplifier gain required

N_{IF} = No. of IF stages required

TABLE XVII

MODIFICATION #6 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
Single Diode Mixer	3.70	2.60	2.09	1.84
(Same as MOD #1)				
Local Oscillator, IF Amplifier, Labor, Misc., Assembly & Test				
(Same as B-O)	18.66	9.86	8.47	6.37
Antenna Unit, Total Costs	22.36	12.46	10.56	8.21
Indoor Unit Total Costs				
(Same as B-O)	19.26	15.73	13.61	13.24
Total Costs, Mod #6	41.62	28.19	24.17	21.45

TABLE XVIII

MODIFICATION #7 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

This table represents the cost savings effected by the removal of a single IF amplifier stage. The results are applicable to other modification costs which are based on a seven stage IF amplifier. If two IF amplifier stages are eliminated, subtract twice the values indicated here.

Parts Removed	10^3	10^4	10^5	10^6
Transistor 2N4996	.21	.15	.10	.10
R.L.C's	.31	.30	.29	.29
Labor and All Other	.20	.15	.11	.10
Total Savings Per Stage	.72	.60	.50	.49

TABLE XIX

MODIFICATION #8 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
Balanced Mixer				
Same as B-"zero"	4.42	3.21	2.60	2.22
Local Oscillator				
Same as B-"zero"	10.02	4.52	4.27	2.27
IF Amplifier				
Substitute one 3N201 for the first IF amplifier	2.96	2.24	1.73	1.73
Labor, Misc, Assembly and Test				
Same as B-"zero"	6.68	3.70	2.82	2.72
Antenna Unit Total Costs	24.08	13.67	11.42	8.94
Indoor Unit Total Costs				
Same as B- "zero"	19.26	15.73	13.61	13.24
Total Costs Mod #8	43.34	29.40	25.03	22.18

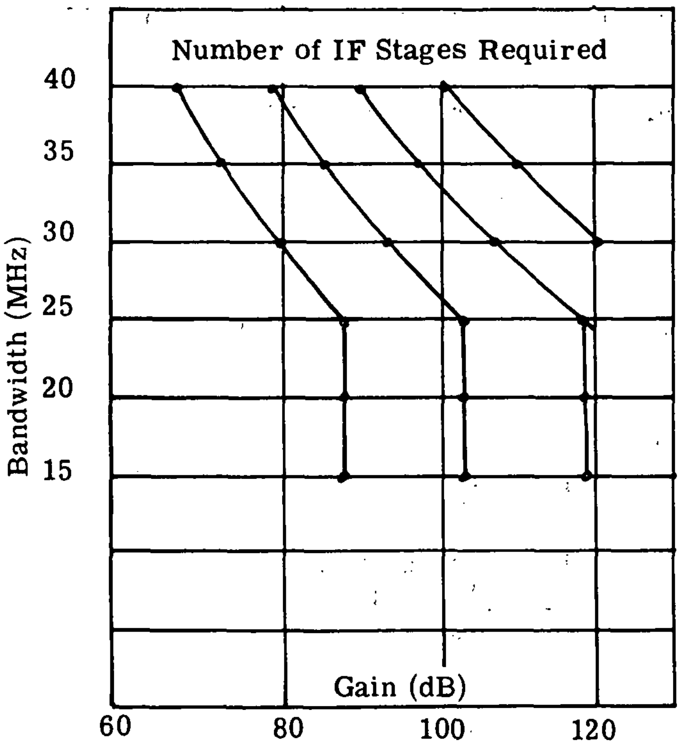
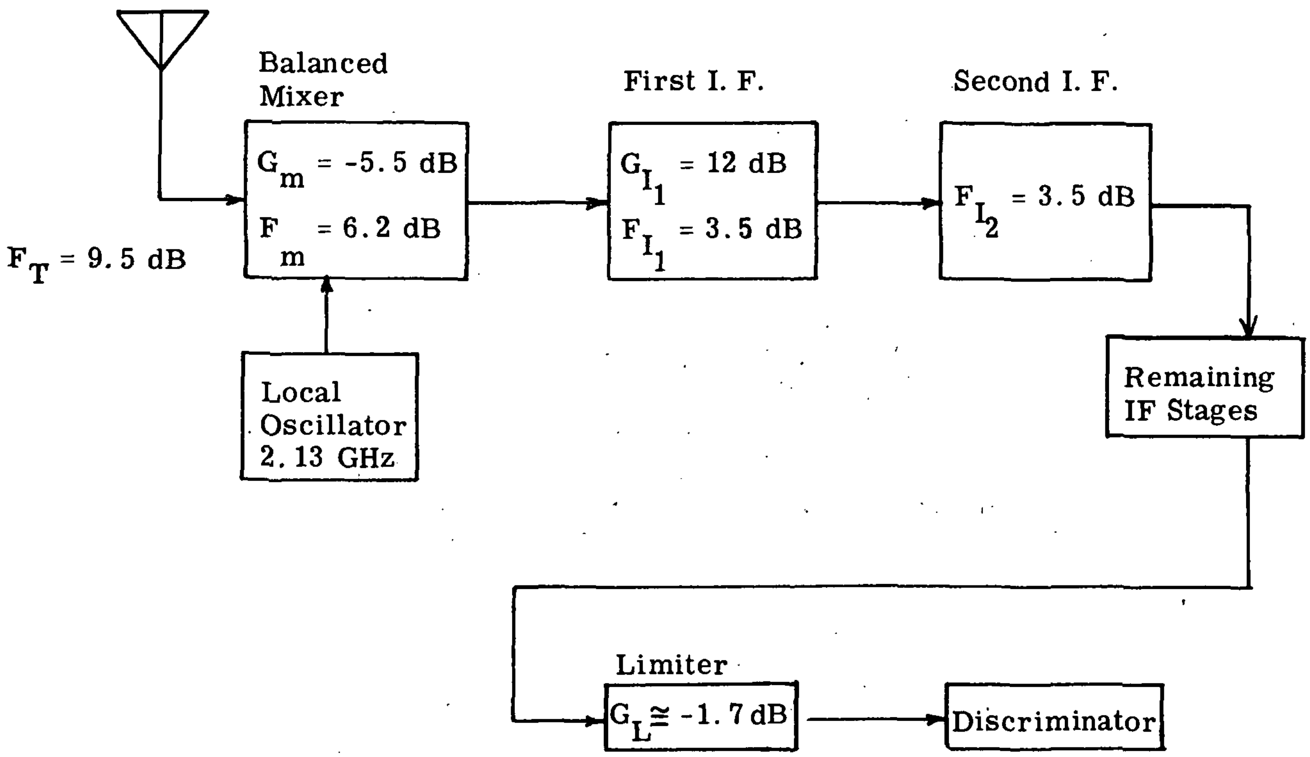


Figure 38. Available IF Gain versus Bandwidth with Number of Stages as Parameter



G_m = Balanced Mixer Conversion Gain

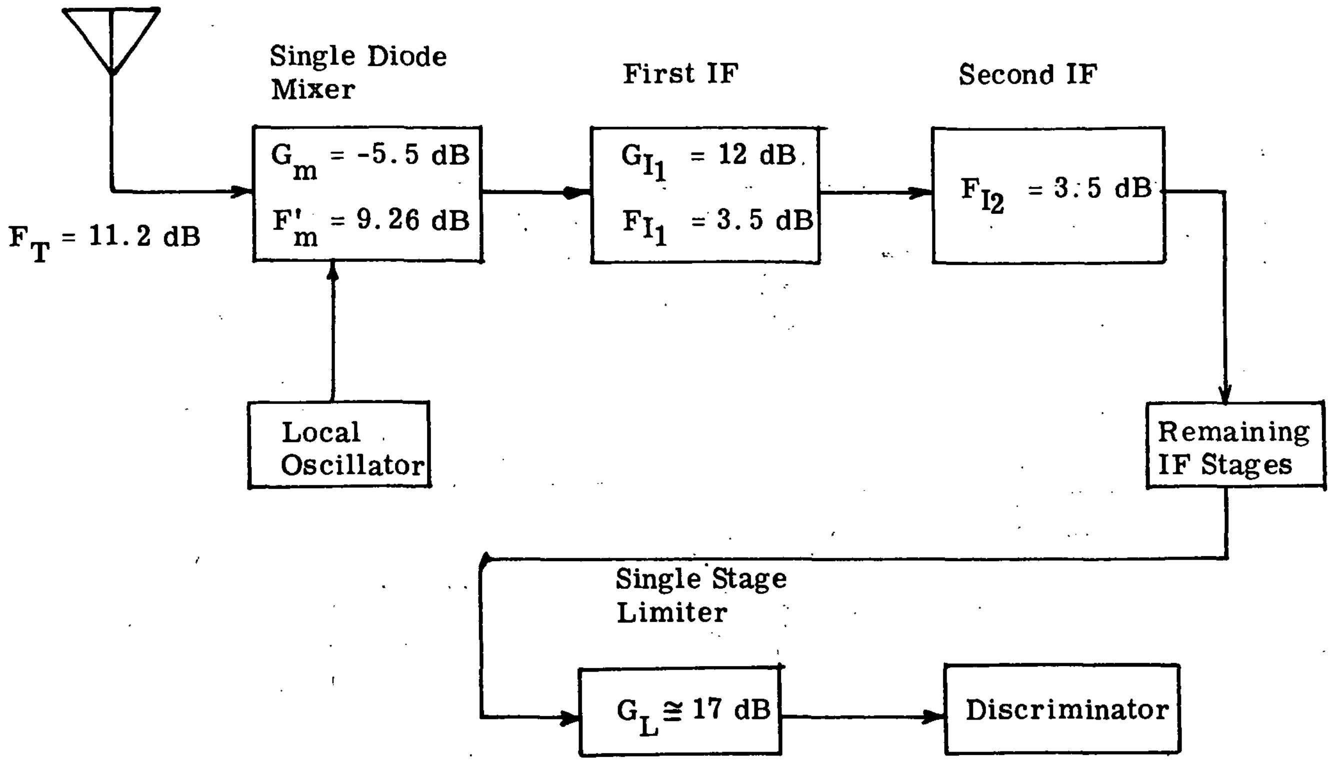
F_m = Balanced Mixer Noise Figure

G_{I_1} = Antenna Unit First I. F. Amplifier Gain

F_{I_i} = Antenna Unit I. F. Amplifier Noise Figure

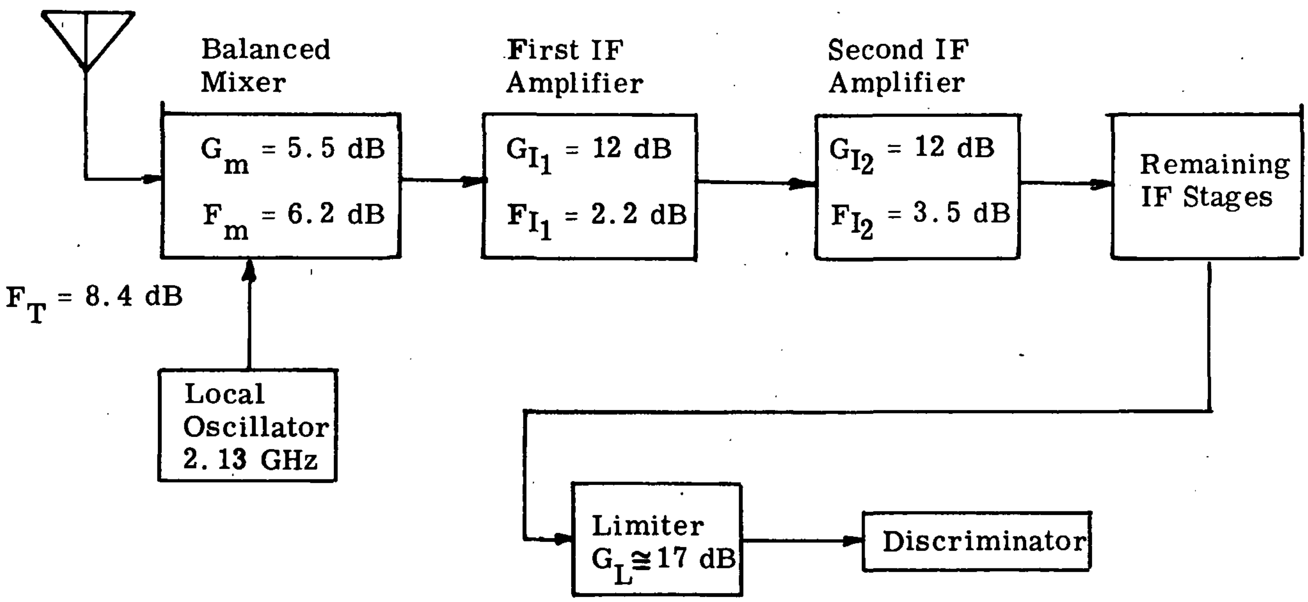
G_L = Limiter Gain

Figure 39. B-"zero" Simplified Block Diagram



G_m = Single Diode Mixer Conversion Gain
F'_m = Single Diode Mixer Noise Figure
G_{I_1} = Antenna Unit First I. F. Amplifier Gain
F_{I_i} = Antenna Unit I. F. Amplifier Noise Figure
G_L = Limiter Gain

Figure 40. Modification #6 Simplified Block Diagram



G_m = Balanced Mixer Conversion Gain

F_m = Balanced Mixer Noise Figure

G_{I_i} = IF Amplifier Gain

F_{I_i} = IF Amplifier Noise Figure

G_L = Limiter Gain

Figure 41. Modification #8 Simplified Block Diagram

Modification #9 Addition of a Transistor RF Preamplifier

Modification #9 is identical to Modification #4 for the S-Band AM converter systems. A system noise figure of 7.2 dB is obtained. The block diagram for this modification is shown in Figure 42, the cost analysis, in Table XX.

Modification #10 Addition of a Parametric RF Preamplifier

Modification #10 is similar to Modification #5 with one exception. The FM system is less susceptible to incidental FM, from the pump or local oscillator sources, than is the AM converter system. This fact is a result of the wide difference between the AM system aural subcarrier deviation and the wide deviation of the FM system carrier. For Modification #10 the single cavity sources for the pump and local oscillator will be adequate. The costs of these sources are assumed to be the same as for the Type D converter local oscillator. A block diagram for Modification #10 is shown in Figure 43. The cost analysis for this modification is contained in Table XXI. The system noise figure with this modification is 2.6 dB.

Further Noise Figure Reduction

The B-21 set of specifications result in a C/N estimate of 11.2 dB at the discriminator when the parametric preamplifier is used. Any further improvement in the converter noise figure would be impractical from a cost viewpoint.

5.3 TYPE D CONVERTER SYSTEMS (12.0 GHz FM)

The base design, D-"zero" simplified block diagram is shown in Figure 44 along with the key performance parameters. A noise figure of 11.25 dB represents the average measured value for the D-"zero" converters that were built.

The Type D converter system modifications are summarized in Tables XXII and XXIII. The procedures used to match the designs to the performance parameters are the same as with the Type B FM systems. The noise figures computed for the Type D modifications are different than those obtained for the lower frequency FM converters.

Modification #11 Substitution of a Single Diode Mixer for the Balanced Mixer in the Type D Converter

In this modification a single diode X-band mixer drives an IF amplifier identical to that found in D-"zero". It is assumed that a single diode mixer can yield the same conversion gain as a balanced mixer, but at a noise figure 3 dB higher. The noise figure resulting from this modification is 13.0 dB. A simplified block diagram of Modification #11 is shown in Figure 45. together with key performance parameters. Table XXIV provides the cost estimate data for this modification.

TABLE XX

MODIFICATION #9 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
R. F. Amplifier				
(Same as Mod #4)	19.52	10.15	8.91	6.80
Remaining Costs				
(Same as B-O)	42.34	28.80	24.68	21.83
Total Costs, Mod #9	61.86	38.95	33.59	28.63

TABLE XXI

MODIFICATION #10 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
X-band Pump				
(Same as D-"zero" L.O.)	39.83	26.07	11.04	5.61
Paramp				
Microwave Associates (MA-4536)	105.00	95.00	67.00	42.50
16 In^2 PPO	2.00	1.26	1.14	1.14
X-Band Mixer				
(Same as D-"zero" Mixer)	13.72	11.34	10.43	8.81
X-Band Local Osc.				
(Same as D-"zero" L.O.)	39.83	26.07	11.04	5.61
IF Amplifier				
(Same as D-"zero" I.F.)	1.99	1.66	1.40	1.40
Misc. Parts				
Enclosure	1.20	.40	.40	.40
Hardware	2.25	1.87	1.50	1.50
Freight and Spoilage	4.40	2.75	2.26	2.58
Total Parts	210.22	166.42	116.21	69.55
Labor, Assy & Test	8.24	4.12	2.76	2.58
Antenna Unit Total Cost	218.46	170.54	118.97	72.13

TABLE XXI (Cont.)

Indoor Unit	10^3	10^4	10^5	10^6
Power Supply	4.06	3.83	3.81	3.81
IF Amplifier, Limiter and Discriminator (Same as B-"zero)	5.45	4.59	3.86	3.60
Video Amplifier-Mod. (Same as B-"zero")	2.99	2.38	2.02	2.02
Miscellaneous				
(Same as B-"zero")	3.52	2.75	2.72	2.61
Total Parts	16.02	13.55	12.41	12.04
Labor, Assy & Test	4.12	2.84	2.06	2.06
Total Indoor Unit Costs	20.14	16.39	14.47	14.10
Total Antenna Unit Costs	218.43	170.52	118.95	72.11
Total Mod #10 Costs	238.57	186.91	133.42	86.21

TABLE XXII

FACTORY COST ESTIMATES - TYPE D CONVERTERS

12.0 GHz FM

Code No.	S_o/N_o dB	F_T dB	$I_{FM}-P_N$ dBW	M.I.	BW MHz	P_N dBW	C/N dB	Mod. No.	Annual Production (1970) 10^3	10^4	10^5	10^6
D-1	25	13.0	134.0	0.8	19	-131.1	13.1	11	77.61	54.31	35.69	28.37
2	26	13.0	135.0	0.9	20	-131.0	13.0	11	77.61	54.31	35.69	28.37
3	27	13.0	136.0	1.0	21	-130.7	12.7	11	77.61	54.31	35.69	28.37
4	28	13.0	137.0	1.1	22	-130.5	12.5	11	77.61	54.31	35.69	28.37
5	29	13.0	138.0	1.2	23	-130.3	12.3	11	77.61	54.31	35.69	28.37
6	30	13.0	139.0	1.4	25	-130.0	12.0	11	77.61	54.31	35.69	28.37
7	31	13.0	140.0	1.6	26	-129.8	11.8	11	77.61	54.31	35.69	28.37
8	32	13.0	141.0	1.7	27	-129.6	11.6	11	77.61	54.31	35.69	28.37
9	33	11.2	140.2	1.6	26	-129.8	13.6		83.92	59.75	40.37	32.78
10	34	11.2	141.2	1.8	28	-129.5	13.3		83.92	59.75	40.37	32.78
0	35	11.2	142.2	2.0	30	-129.2	13.0		83.92	59.75	40.37	32.78
11	36	11.2	143.2	2.2	32	-128.9	12.7		83.92	59.75	40.37	32.78
12	37	11.2	144.2	2.4	34	-128.7	12.5		83.92	59.75	40.37	32.78
13	38	11.2	145.2	2.8	37	-128.2	12.0		83.92	59.75	40.37	32.78
14	39	10.2	145.2	2.8	37	-128.2		13	84.89	60.33	40.70	33.11
15	40	7.8	143.8	2.4	34	-128.7	15.9	14	149.49	107.71	76.36	64.56
16	41	7.8	144.8	2.6	35	-128.5	15.7	14	149.49	107.71	76.36	64.56
17	42	7.8	145.8	3.0	39	-128.0	15.2	14	149.49	107.71	76.36	64.56
18	43	4.5	143.5	2.2	32	-128.9	19.4	12	377.87	165.12	119.34	99.21
19	44	4.5	144.5	2.6	35	-128.5	19.0	12	377.87	165.12	119.34	99.21
20	45	4.5	145.5	2.8	37	-128.2	18.7	12	377.87	165.12	119.34	99.21

TABLE XXIII

FACTORY COST ESTIMATES - TYPE D CONVERTERS

12.0 GHz FM

Code No.	P_S dBW	F_T dB	I_{FM}-P_N dBW	M.I.	BW MHz	P_N dBW	C/N dB	G_{IF} dB	N_{IF}	Mod. No.	Annual Production (1970) 10^3	10^4	10^5	10^6
D-21	-114	4.5	144.5	2.6	35	-128.5	10.0**	102	7	12	377.87	165.12	119.34	99.21
22	-113	4.5	143.5	2.2	32	-128.9	11.4**	101	7	12	377.87	165.12	119.34	99.21
23	-112	4.5	142.5	2.0	30	-129.2	12.7	100	7	12	377.87	165.12	119.34	99.21
24	-111	4.5	141.5	1.8	28	-129.5	14.0	99	7	12	377.87	165.12	119.34	99.21
25	-110	4.5	140.5	1.6	26	-129.8	15.3	98	7	12	377.87	165.12	119.34	99.21
26	-109	7.8	143.8	2.4	34	-128.7	11.9	97	7	14	149.49	107.71	76.36	64.56
27	-108	7.8	142.8	2.1	31	-129.1	13.3	96	7	14	149.49	107.71	76.36	64.56
28	-107	10.2	143.2	2.2	32	-128.9	11.7	95	7	13	84.89	60.33	40.70	33.11
29	-106	11.2	143.2	2.2	32	-128.9	11.7	94	7		83.92	59.75	40.37	32.78
0.	-105	11.2	142.2	2.0	30	-129.2	13.0	93	7		83.92	59.75	40.37	32.78
											83.92	59.75	40.37	32.78
30	-104	13.0	143.0	2.2	32	-128.9	11.9	92	6	11	76.89	53.71	35.19	27.88
31	-103	13.0	142.0	2.0	30	-129.2	13.2	91	6	11	76.89	53.71	35.19	27.88
32	-102	13.0	141.0	1.7	27	-129.6	14.6	90	6	11	76.89	53.71	35.19	27.88
33	-101	13.0	140.0	1.6	26	-129.8	15.8	89	6	11	76.89	53.71	35.19	27.88
34	-100	13.0	139.0	1.4	25	-130.0	17.0	88	6	11	76.89	53.71	35.19	27.88
35	- 99	13.0	138.0	1.2	23	-130.3	18.3	87	5	11	76.17	53.11	34.69	27.39
36	- 98	13.0	137.0	1.1	22	-130.5	19.5	86	5	11	76.17	53.11	34.69	27.39
37	- 97	13.0	136.0	1.0	21	-130.7	20.7	85	5	11	76.17	53.11	34.69	27.39
38	- 96	13.0	135.0	0.9	20	-131.0	22.0	84	5	11	76.17	53.11	34.69	27.39
39	- 95	13.0	134.0	0.8	19	-131.1	23.1	83	5	11	76.17	53.11	34.69	27.39
40	- 94	13.0	133.0	0.7	18	-131.4	24.4	82	5	11	76.17	53.11	34.69	27.39
41	- 93	13.0	132.0	0.7	18	-131.4	25.4	81	5	11	76.17	53.11	34.69	27.39
42	- 92	13.0	131.0	0.6	17	-131.6	26.6	80	5	11	76.17	53.11	34.69	27.39
43	- 91	13.0	130.0	0.5	16	-131.8	27.8	79	5	11	76.17	53.11	34.69	27.39
44	- 90	13.0	129.0	0.5	16	-131.8	28.8	78	5	11	76.17	53.11	34.69	27.39

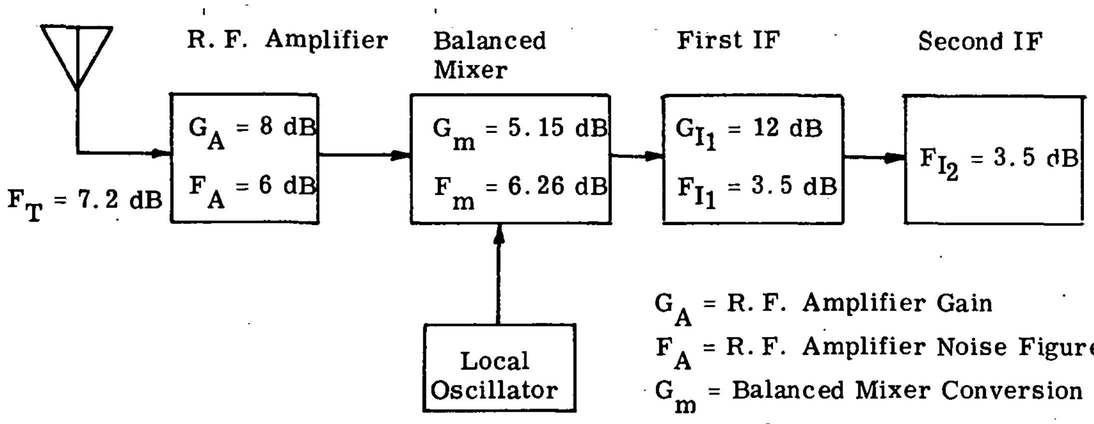
* S_o/N_o = 35 dB

**These converters are low on C/N but next NF improvement is too expensive.

TABLE XXIV

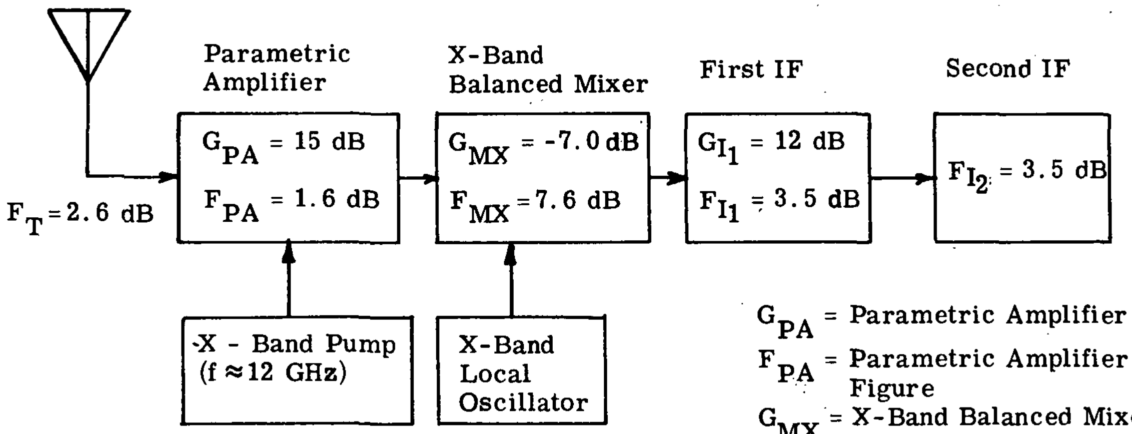
MODIFICATION #11 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
Single Diode Mixer				
1 HP2740 Diode	5.50	4.75	4.50	3.75
Connector	1.72	1.21	.86	.74
5.5 In^2 PPO PCB	.69	.44	.39	.39
	7.91	6.40	5.75	4.88
Local Oscillator				
Same as D-"zero"	39.83	26.07	11.04	5.61
IF Amplifier, Misc, Labor, etc.				
Same as D-"zero"	10.61	6.11	4.78	4.62
Antenna Unit, Total Costs	58.35	38.58	21.57	15.11
Indoor Unit, Total Costs				
(Same as D-"zero")	19.26	15.73	14.12	13.26
Total Cost, Mod #11	77.61	54.31	35.69	28.37



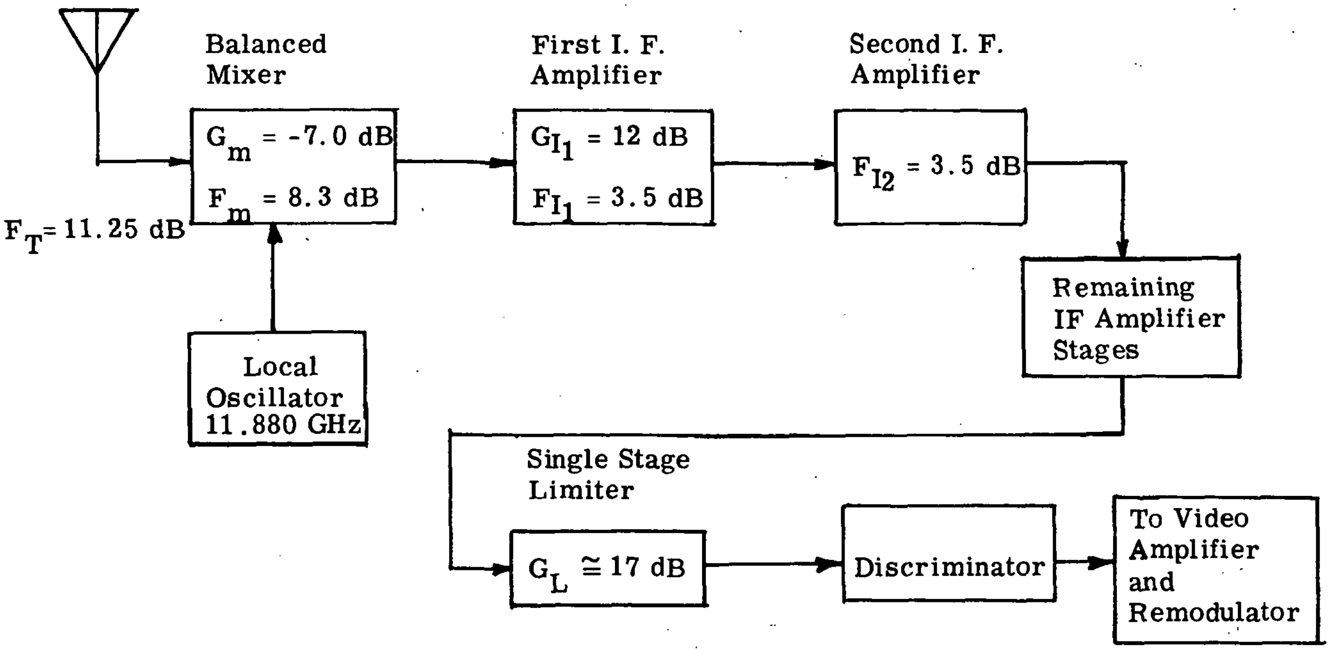
G_A = R. F. Amplifier Gain
F_A = R. F. Amplifier Noise Figure
G_m = Balanced Mixer Conversion Gain
F_m = Balanced Mixer Noise Figure
G_{I_1} = Antenna Unit First I. F. Amplifier Gain
$F_{I_{1,2}}$ = Antenna Unit I. F. Amplifier Noise Figure

Figure 42 . Modification #9 Simplified Block Diagram



G_{PA} = Parametric Amplifier Gain
F_{PA} = Parametric Amplifier Noise Figure
G_{MX} = X-Band Balanced Mixer Conversion Gain
F_{MX} = X-Band Balanced Mixer Noise Figure
G_{I_1} = Antenna Unit First I. F. Amplifier Gain
$F_{I_{1,2}}$ = Antenna Unit I. F. Amplifier Noise Figure

Figure 43. Modification #10 Simplified Block Diagram



G_m = Balanced Mixer Conversion Gain

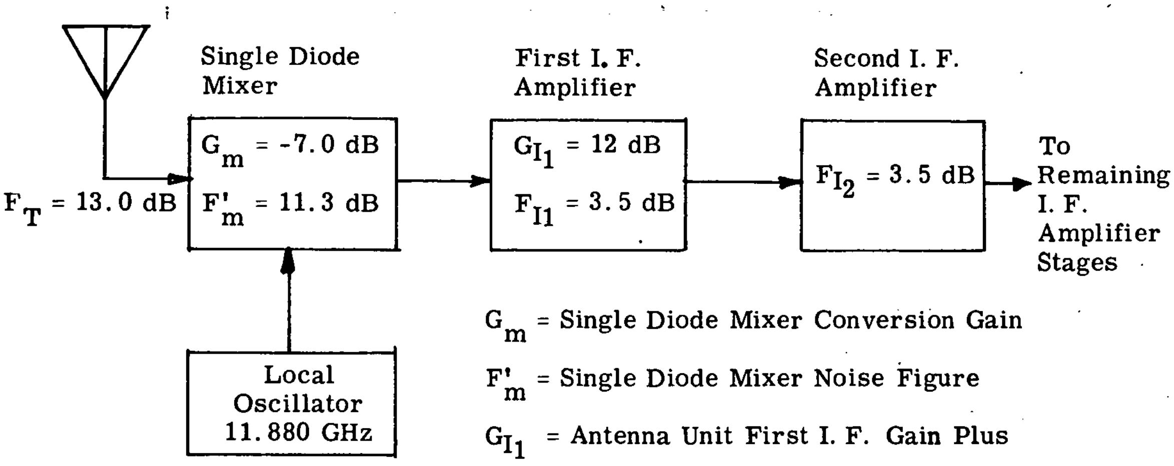
F_m = Balanced Mixer Noise Figure

G_{I_1} = First I. F. Amplifier Gain

$F_{I_{1,2}}$ = I. F. Amplifier Noise Figure

G_L = Limiter Gain

Figure 44. D-zero Simplified Block Diagram



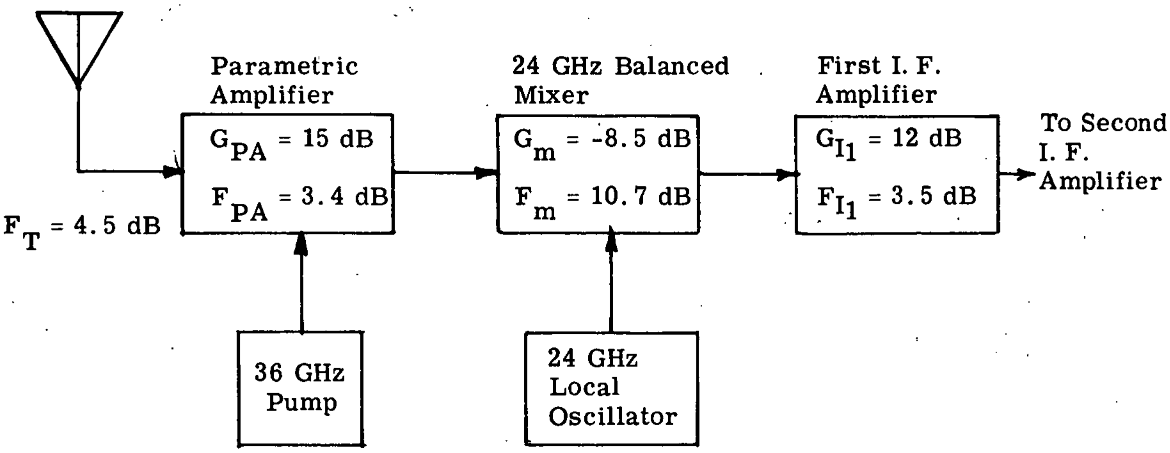
G_m = Single Diode Mixer Conversion Gain

F'_m = Single Diode Mixer Noise Figure

G_{I_1} = Antenna Unit First I. F. Gain Plus

$F_{I_{1,2}}$ = Antenna Unit I. F. Amplifier Noise Figure

Figure 45 . Modification #11 Simplified Block Diagram



G_{PA} = Parametric Amplifier Gain

F_{PA} = Parametric Amplifier Noise Figure

G_m = Mixer Conversion Gain

F_m = Mixer Noise Figure

G_{I1} = Antenna Unit First I. F. Amplifier Gain

F_{I_1} = Antenna Unit I. F. Amplifier Noise Figure

Figure 46. Modification #12 Simplified Block Diagram

Modification #12 <u>Addition of a Parametric RF Amplifier to the Type D Converter</u>

Figure 46 is a simplified block diagram of Modification #12 and shows an X-Band parametric amplifier-balanced mixer system driving an IF amplifier identical to that in D-"zero". The noise figure and gains of the various block are estimated values. The pump oscillator is of the Gunn diode type, operating at 36 GHz, and the local oscillator, also a Gunn oscillator, operates at about 24 GHz.

This modification will provide a noise figure of 4.5 dB. The cost analysis is shown in Table XXV.

<u>Modification #13 Improved IF Amplifier for the Type D Converter</u>

Modification #13 is similar to Modification #8. When applied to the Type D converter a system noise figure of 10.2 dB results. The block diagram for this modification is shown in Figure 47 along with the performance parameters. The cost analysis for this modification is presented in Table XXVI.

<u>Modification #14 Type D Converter with Tunnel Diode Preamplifier</u>

This modification uses a tunnel diode preamplifier with a gain of 10 dB and a noise figure of 7 dB to achieve a system noise figure of 7.8 dB as shown in the block diagram, Figure 48. The incremental cost is detailed in Table XXVII.

TABLE XXV

MODIFICATION #12 DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1970)

Antenna Unit	10^3	10^4	10^5	10^6
36 GHz Pump				
Varian VSA9210 Gunn Diode	147.00	50.00	34.00	26.00
Cavity, Transistors, R, L, C (Same as D-"zero" L.O.)	6.83	4.07	3.04	2.86
	153.83	54.07	37.04	28.86
24 GHz Pump				
Varian VSK9204 Gunn Diode	125.00	42.50	28.98	22.10
Cavity, Transistors, R, L, C (Same as 36 GHz Pump)	6.83	4.07	3.04	2.86
	131.83	46.57	32.02	24.96
Varactor Diode Amp				
Varian VAP-104 Varactor Diode	39.10	25.30	16.10	10.20
16 In^2 PPO PCB	2.00	1.26	1.14	1.14
	41.10	26.56	17.24	11.34
Balanced Mixer				
(Similar to D-"zero" Mixer)	13.72	11.34	10.43	8.81
IF Pre Amp				
(Same as D-"zero")	1.99	1.66	1.40	1.40
	342.47	140.20	98.13	75.37
Misc. Parts				
(Includes Freight & Spoilage)	7.35	4.55	3.80	3.50
Labor, Assembly & Test	8.24	4.12	2.77	2.58
Antenna Unit Total Costs	358.06	148.87	104.70	81.45

TABLE XXV (cont.)

Indoor Unit	10^3	10^4	10^5	10^6
Power Supply	4.24	4.00	3.98	3.98
IF Amplifier, Limiter Discriminator, Video Amp, Modulator, Misc., Labor, Assembly & Test				
(Same as D-"zero")	15.57	12.25	10.66	9.78
Indoor Unit Total Costs	19.81	16.25	14.64	17.76
Total Costs, Mod #12	377.87	165.12	119.34	99.21

TABLE XXVI

MODIFICATION #13 IMPROVED IF AMPLIFIER
FOR THE TYPE D CONVERTER

Antenna Unit	10^3	10^4	10^5	10^6
Balanced Mixer (Same as D-"zero")	13.72	11.34	10.43	8.81
Local Oscillator (Same as D-"zero")	39.83	26.07	11.04	5.61
IF Amplifier One 3N201 in place of 1 2N4996	2.96	2.24	1.73	1.73
Misc. Parts (Same as D-"zero")	4.50	2.39	2.00	1.93
Total Parts	61.01	42.04	25.20	18.08
Labor, Assy. & Test	4.12	2.06	1.38	1.29
Antenna Unit Total Cost	65.13	44.10	26.58	19.37
Indoor Unit Total Cost (Same as D-"zero")	19.76	16.23	14.12	13.74
Total Costs, Mod. #13	84.89	60.33	40.70	33.11

TABLE XXVII

MODIFICATION # 14 TYPE D CONVERTER WITH TUNNEL DIODE PREAMPLIFIER

	10^3	10^4	10^5	10^6
Mod # 13 D Converter	84.89	60.33	40.70	33.11
Tunnel Diode Preamplifier				
Tunnel Diode	40.00	30.00	22.00	18.50
Circulator SS	.62	.11	.05	.04
Resistors (50Ω)	3.00	3.00	3.00	3.00
Ferrite	1.14	1.10	1.10	1.10
Magnets	.24	.16	.12	.11
Connectors (2)	4.20	2.90	2.16	1.82
Enclosure	.79	.55	.41	.35
Misc.	1.50	1.20	1.00	.90
Freight & Spoilage	1.52	.98	.73	.66
Materials	53.01	40.00	30.57	26.48
Processing	8.59	5.40	3.26	3.14
Assy. & Test	3.00	1.98	1.83	1.83
Total Amplifier Cost	64.60	47.38	35.66	31.45
Total Modification # 14 Cost	149.49	107.71	76.36	64.56

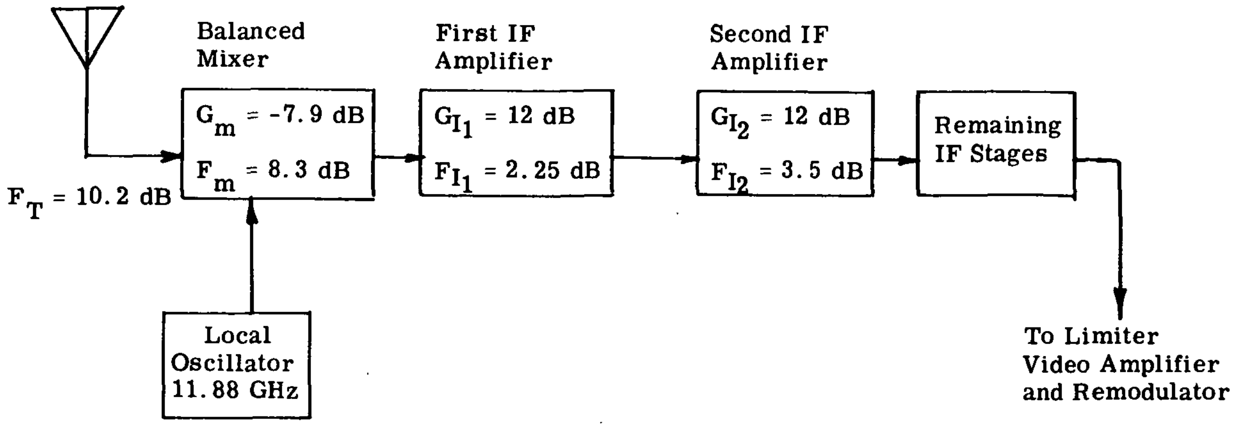


Figure 47 . Type D Converter Simplified Block Diagram with Improved IF Amplifier

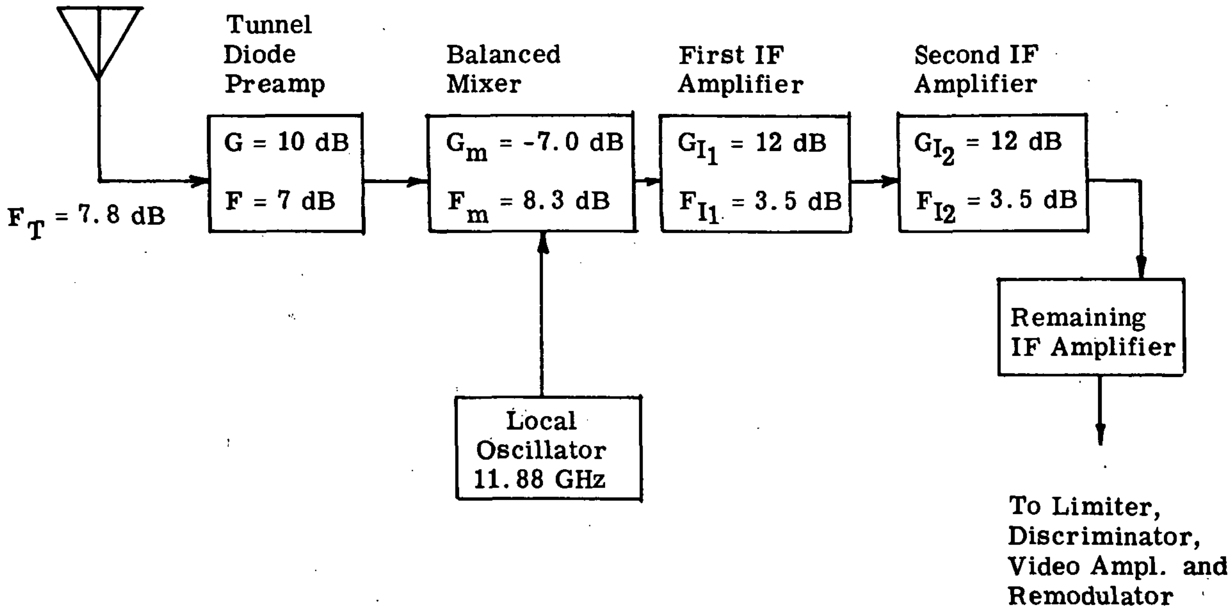


Figure 48. Modification #14 Simplified Block Diagram

6.0 INTEGRATED TUNNEL DIODE AMPLIFIER

6.1 OBJECTIVES

The purpose of this task is to analyze design, fabricate and test a low-cost tunnel diode preamplifier with microstrip circulators, to be used as part of a ground receiving system under the following constraints:

Frequency	12 GHz
Bandwidth	240 MHz
Noise Figure	7 dB max.
Gain	10 dB min.
Intermodulation Products	-40 dB max. at -45 dBM signal input
Temperature Range	-40°F to plus 130 degree F
Power Supply	12 volts dc
RF Connectors	compatible with 12 GHz ground signal processor

6.2 SUMMARY

An integrated tunnel diode amplifier structure operating at 12 GHz constitutes a new venture, the success of which depends heavily on very careful processing and thoughtful design. To ease the design and fabrication schedule, it was decided that the circulator, the tunnel diode and stabilizing components, and the bias circuitry be located on three separate substrates.

Circulator designs, using both an alumina-garnet structure and an all-garnet structure were evaluated; both designs yield fairly good results.

The tunnel diode fabricated on a glass substrate did not go too well. The lengthy sequence of 1) preparing the germanium-glass billet, 2) the metallization, 3) gold plating, 4) drilling, 5) diode junction formation and 6) attachment of the pill resistor, comprises a delicate and time-consuming operation. The overall yield was small, with the greatest losses occurring at the junction formation stage. The consequence was that our effort to meet the goals was seriously degraded.

In the course of testing and design one important fact, which eluded the designer's attention and caused an unwarranted degree of optimism, was the marginal stability. The unit appeared to be capable of yielding a 10 to 12 dB gain over a 100 MHz frequency range and no steady oscillation was observed. However, it was later found that this gain was, actually, signal-level dependent. For example, at a -45 dBm input the signal unit gives an 11 dB gain; but it gives only a 6 dB gain at lower signal levels. And also, because of marginal stability, the dc bias is critical to adjust.

Some hope of eliminating these difficulties was demonstrated when matching adjustments between the tunnel diode substrate and the circulator were introduced. The amplifier gain and bandwidth, with legitimate signal levels, were excellent (See Figure 49); but later the diode was damaged when the mechanical assembly was being fixed.

Unfortunately, this observation was made too late in the program; there were not enough funds to carry out the necessary corrective measures and redesign. Nevertheless, continuous effort has been made, although at a slow pace, to find the adverse causes and correct these. A modified design has been completed and is pending assembly with discrete tunnel diodes. This additional effort is presented in Section 6.10, Addendum.

In conclusion, a TDA that meets all of the specifications is feasible. The cost of a totally integrated structure will be more expensive than a partially integrated structure using discrete devices, because of the yield problem. The integration of the TDA and mixer in a compatible structure is a desirable objective.

6.3 FUNCTIONAL BLOCKS OF THE UNIT AND CIRCUIT CONSIDERATIONS

The functional block diagram of the tunnel diode amplifier is presented in Figure 50. The essential parts are the four port circulator, the TD with its stabilization circuits, and the dc voltage regulator. The filter is optional.

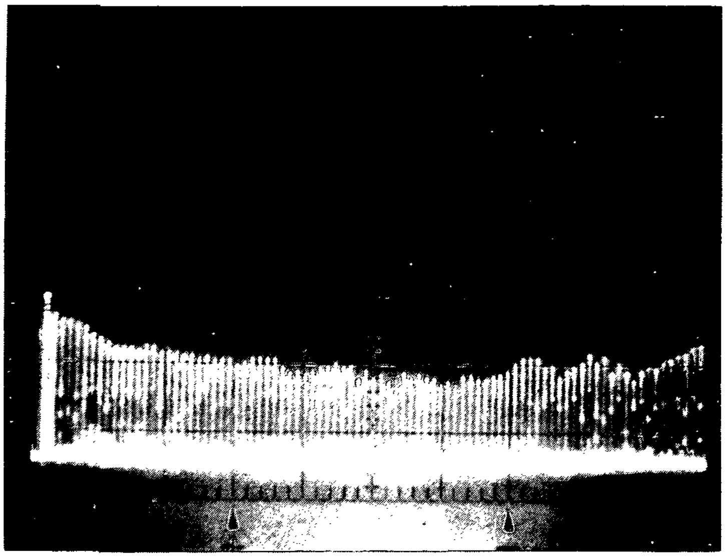
The circulator serves two functions; 1) to provide directivity so that a reflective amplifier incorporating a two-terminal tunnel diode device is possible and 2) to isolate the local oscillator power from the tunnel diode. Low forward loss, high isolation in the reverse direction, and a low VSWR of the ports, are the requirements of the circulator. The tunnel diode circuit must be stable and have a high cutoff frequency to provide gain at the frequencies of interest. The dc regulator eliminates temperature drift and voltage fluctuations and provides means for adjusting the initial setting.

The constraints that 1) the dc supply polarity be positive, 2) the tunnel diode semiconductor be made of p-type material, and 3) the input of the mixer be dc shorted, leads to the arrangement shown in Figure 51, which allows for a dc bias and rf compatibility between the TDA and the mixer.

6.4 CIRCULATOR DESIGN

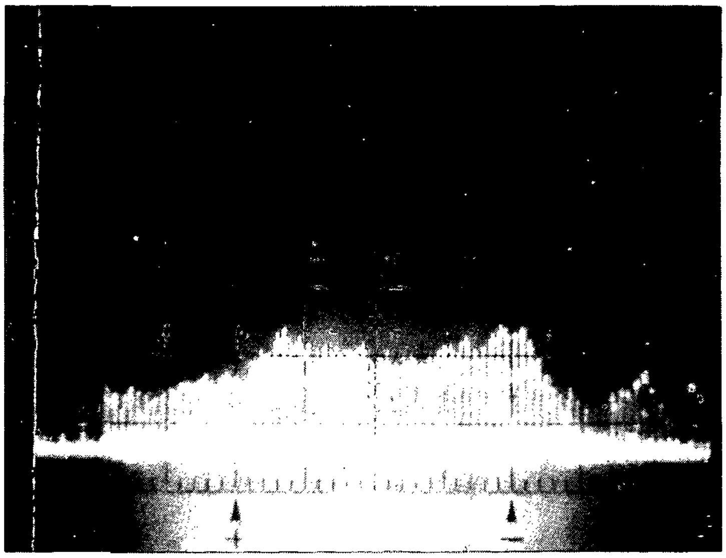
The basic principles of circulator design, using ferrites or garnet materials in strip or microstrip configurations, are well understood. * However, the available garnet or ferrite materials are, in general, not specified for the frequencies of interest here. It is, therefore, a task of the designer to choose and evaluate the properties of those materials that show promise for this application.

* Fay and Comstock, MTT-13, Jan. 1965, p. 15-27. (for example)



Vertical 10 dB/cm.
Horizontal 30 MHz/cm.

(a)



Vertical 10 dB/cm.
Horizontal 100 MHz/cm.

(b)

Figure 49. TDA Gain with Matching,
(approx. 10 dB in the center band)

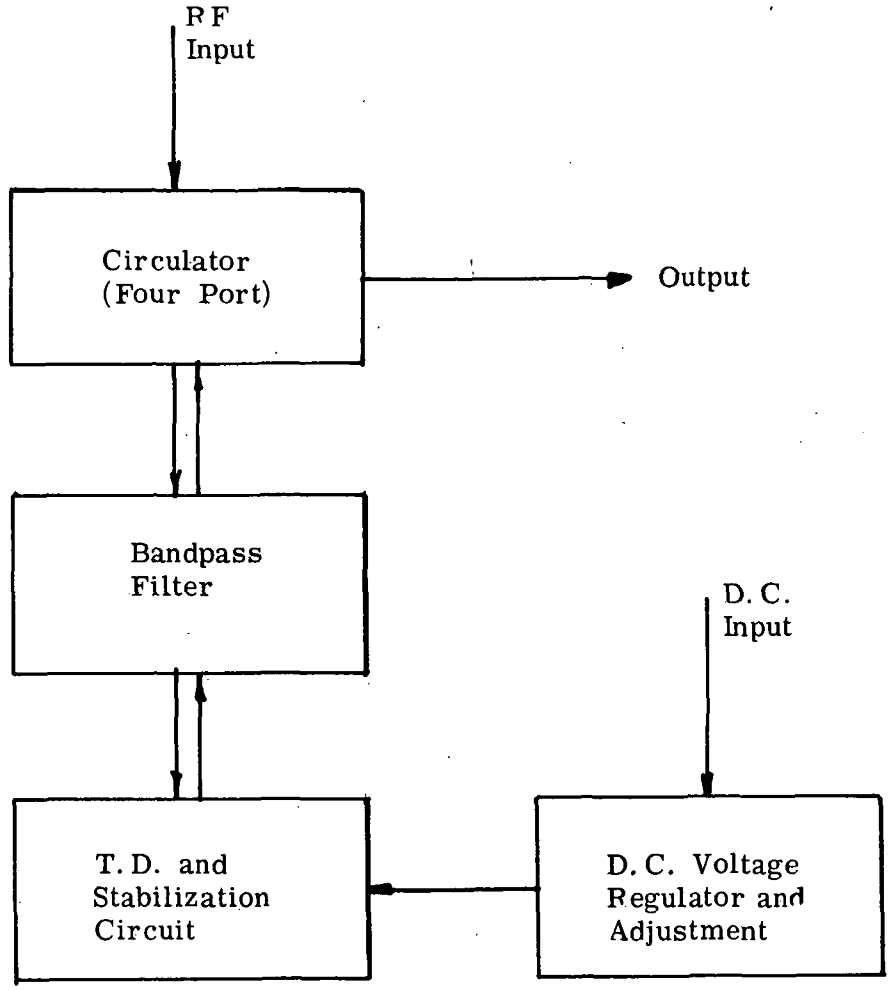


Figure 50. Block Diagram of the T.D.A.

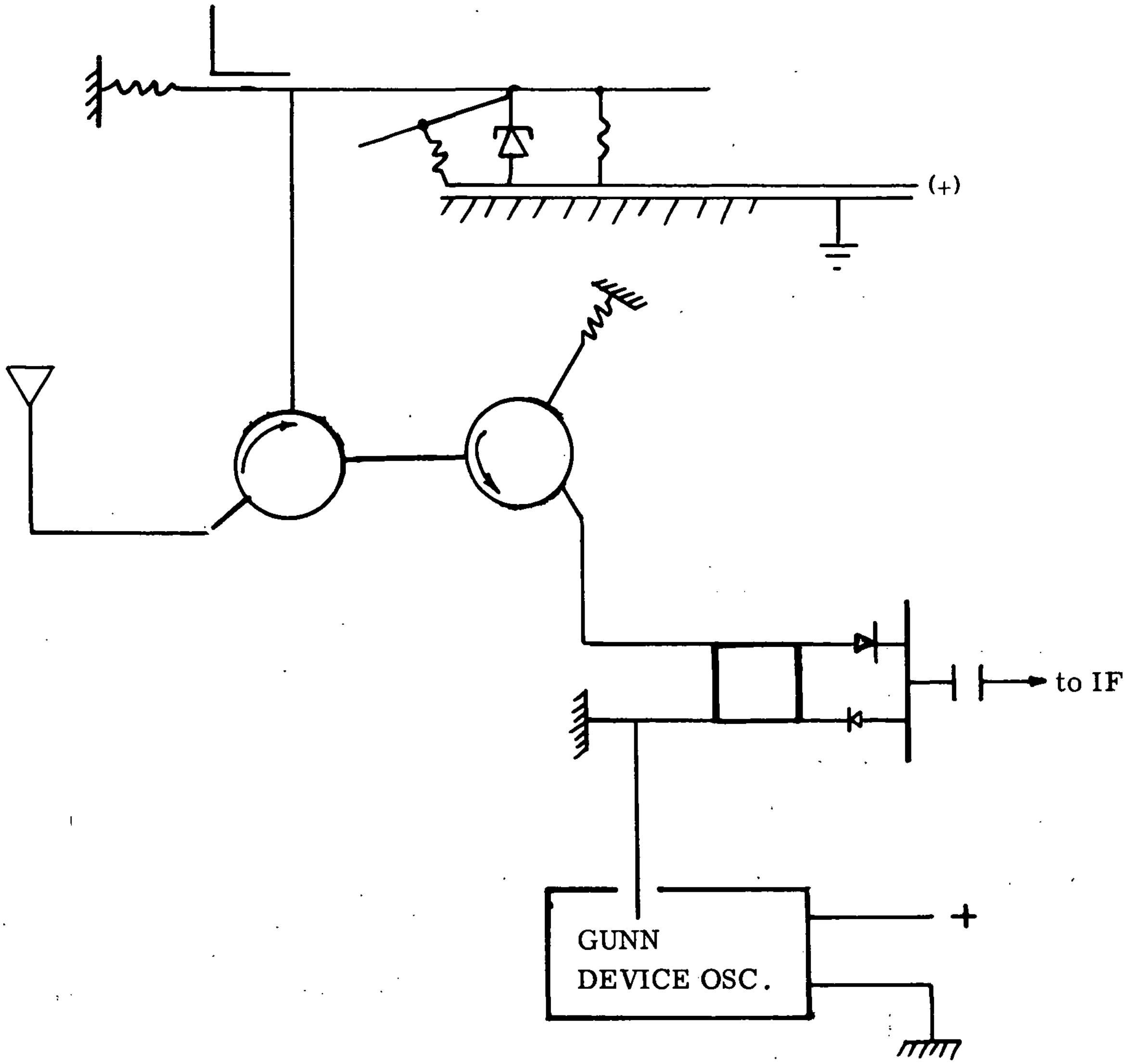


Figure 51. DC Bias and AC Compatibility of TDA and Mixer

Fortunately, a material readily available from TRANSTECH (Type G1001 garnet) is satisfactory, for both mechanical and electrical requirements. G1001 is gadolinium doped and is known to be suitable for higher power applications; also it has good thermal stability characteristics. From known properties of the similar materials the effective change of propagation velocity over the -40 to 55^{0}C range is about 3.5%. The effect on the circulator is expected to be minor. The pertinent properties of garnet and alumina are compared below.

Properties	Ferrite (garnet)	Alumina 99.5%
ϵ'_r	15.8	9.7
$\tan\ \delta_e + \tan\ \delta_\mu$	1×10^{-3}	4×10^{-4}
Q_0	255	258
loss/wave length (conductor)	0.11 dB	0.11 dB

It seems that, to achieve low loss, the best combination of these materials would be a garnet disc and an alumina substrate. The thermal expansion of garnet, alumina, glass and epoxy are:

Material	Coefficient of Thermal Expansion
Alumina (99.7%)	$5 \times 10^{-6}/^{o}C$
glass K650	5×10^{-6}
Ferrite	9×10^{-6}
Epoxy	35×10^{-6}

There are obvious mismatches in the expansion coefficients of these materials. The expansion coefficient of epoxy is too large; a more elastic material, e.g., FEP* would be better as a bonding material. Keeping this as an alternative, conventional techniques using epoxy are still used if the dimensions are small enough so that no excessive stresses will be involved.

The electrical design is based on the following calculations and assumptions:

The substrates are 0.025" Al_2O_3 (99.5%) and 0.025" TRANSTECH Garnet, G-1001, which has the following properties.

- Saturation magnetization ($4\pi M_s$) = 1200 gauss
- g-effective 1.99
- dielectric constant, ϵ 15.2
- gyromagnetic ratio, γ 2.8 MHz/Oersted

*FEP TEFLON (Fluorinated Ethylene Propylene).

Then the effective propagation constant, k, is:

$$k^2 = \omega^2 \epsilon \epsilon_o \frac{\mu^2 - \kappa^2}{\mu} \mu_o$$

where μ and k are the components of the tensorial permeability, $[\mu]$, in $B = \mu H$.

$$[\mu] = \begin{bmatrix} \mu & 0 & -j\kappa \\ 0 & \mu & 0 \\ j\kappa & 0 & \mu \end{bmatrix}$$

The expression, $(\mu^2 - \kappa^2)/\mu$, can be considered as the effective permeability o. the material. It has been shown that μ and κ are related to the physical constants and frequencies by:

$$\mu = 1 + \sigma \kappa$$

$$\kappa = \frac{P}{\sigma^2 - 1}$$

where $$= \frac{\gamma H_i}{f}$$

$$P = \frac{\gamma 4\pi M_s}{f}$$

H_i = internal magnetic field.

A substantial simplification can be made by assuming that $H_i = 0$; i.e., the externally applied magnetization is equal to the saturation magnetization of the material. Thus we have $\sigma = 0$ and $\mu/\kappa = -1/P$. Now

$$P = \frac{2.8 \times 10^6 \times 1200}{12 \times 10^9} = 0.280$$

$$\mu/\kappa = -1/0.280$$

and

$$\mu_{effective} = \frac{\mu^2 - \kappa^2}{\mu} = 0.9219.$$

Applying a correction for the filling factor for the microstrip lines gives:

$$1/\mu' = q/\mu_{eff} + (1-q) = 0.8/0.9219 + 0.2 = 1.068$$

$$\mu' = 0.925$$

Similarly,

$$\epsilon' = 1 + 0.8\,(15.2 - 1) = 12.4 \quad .$$

Substituting these values in the expression for the propagation constant, k.

$$k^2 = (2\pi)^2\,(12 \times 10^9)^2 \times 12.4 \times 0.9219 \times 1/(3 \times 10^{10})^2,$$

and

$$k = 8.5 \text{ radian/cm.}$$

For a cylindrical cavity, resonance mode 110, the relation

$$kR = 1.84$$

determines the radius, R. Hence

$$R = 0.2165 \text{ cm} = 0.0853''$$

To find the matching section, the following procedure was followed. The effective admittance is defined as

$$Y_{eff} = -\sqrt{\frac{\epsilon_o \epsilon'}{\mu_o \mu}} = \frac{1}{377}\sqrt{\frac{12.4}{0.9219}} = 0.00975$$

Then the conductance of the input to the cavity is given by

$$G_R = \frac{4R}{d}\,Y_{eff}\left|\frac{\kappa}{\mu}\right|$$

where d is the thickness of the ferrite disc.

$$G_R = \frac{4 \times 0.0853}{0.025} \times 0.00975 \times 0.280 = 0.0372 \text{ mhos}$$

or

$$Z = 26.9 \text{ ohms}$$

To be matched to 50 ohms by a quarter wave length section, the matching section must have $Z_o = 36.6$ ohms.

These values were used for our initial design. Subsequently changes were made to zero in on the frequency characteristics. The disc radius was reduced to 0.071" and the matching line width to 0.037" (see Figure 52). The same dimensions were used for the layout of the design shown in Figure 53, which included a filter and an arm extended to the TD substrate. The measured characteristics are shown in Tables XXVIII through XXXI. The forward losses and isolation characteristics are plotted in Figure 54. The loss through the four-port circulator in the forward direction is, therefore, 3 dB, including the line and connector losses. The isolation from output port to the TD port is better than 44 dB at the oscillator frequency of 11.88 GHz. The input of the circulator has a VSWR of about 1.2, and is slightly capacitive. When looking from the TD port, the impedance is nearly resistive and smaller than 50 ohms. This value is, indeed, very critical to the design and will be discussed in a later section.

The thermal effect over a range of -40 to +55°C is shown in Figure 55. These characteristics are not beyond expectations and will not cause any serious deterioration of function.

The permanent magnets used are cut from Indox 7 material (Indiana General); the dimensions are 0.2" × 0.2" × 0.4", located two above and two below. Indox 7 has a B_r of about 2500 gauss and is very adequate for this application.

The effect of packaging on the circulator characteristics was observed. It was concluded that the cover plate should be thinned, to allow as much air between the substrate and cover as possible. This revision is not indicated on the drawings.

6.5 FILTER DESIGN

The rejector filter design and its characteristics are described here.

The functions of the rejector filter is to provide a shunt branch that presents a resistive load of about 50 ohms over all regions except in and near the band of amplification. The coupled resonator, as shown in Figure 56 fits this demand. The reflection and transmission characteristics are shown in the photos, Figures 57a and 57b. In the pass-band of the amplifier, the input of the filter is about 240 - j 150 ohms. When this impedance is shunted across a 50 ohm line, the loss will be of the order of 0.5 dB. This shunt branch, in combination with the stabilizing circuit on the diode substrate, will therefore insure the absolute stability of the amplifier.

While this filter effectively prevented instability at a frequency far removed from the operating range, our later experience cast some doubts on its usefulness in the immediate neighborhood (± 1 GHz) of 12 GHz. The reactance associated with the filter may complicate the situation, or lessen the filter's effectiveness as a stabilizer. Therefore, it is more appropriate to consider this component as optional.

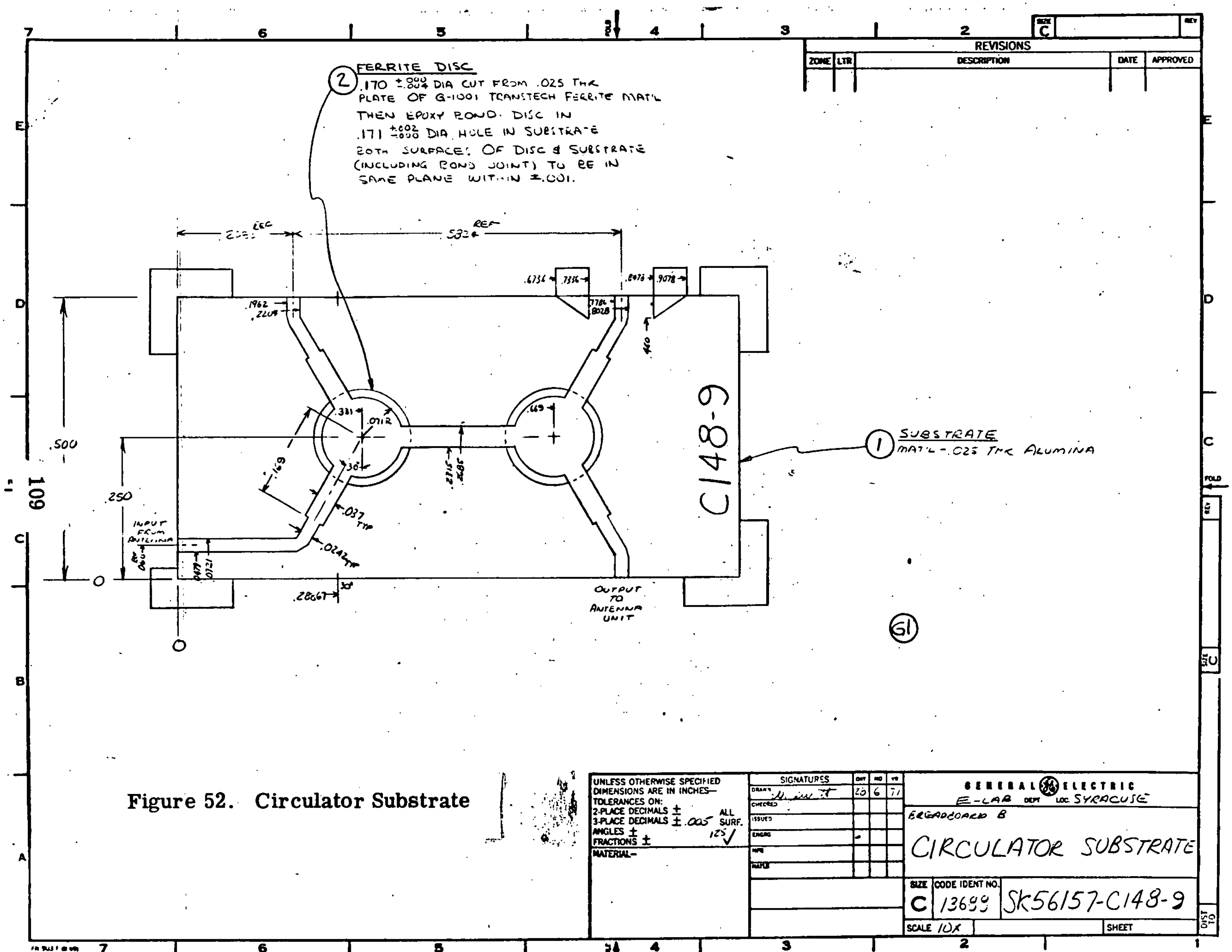


Figure 52. Circulator Substrate

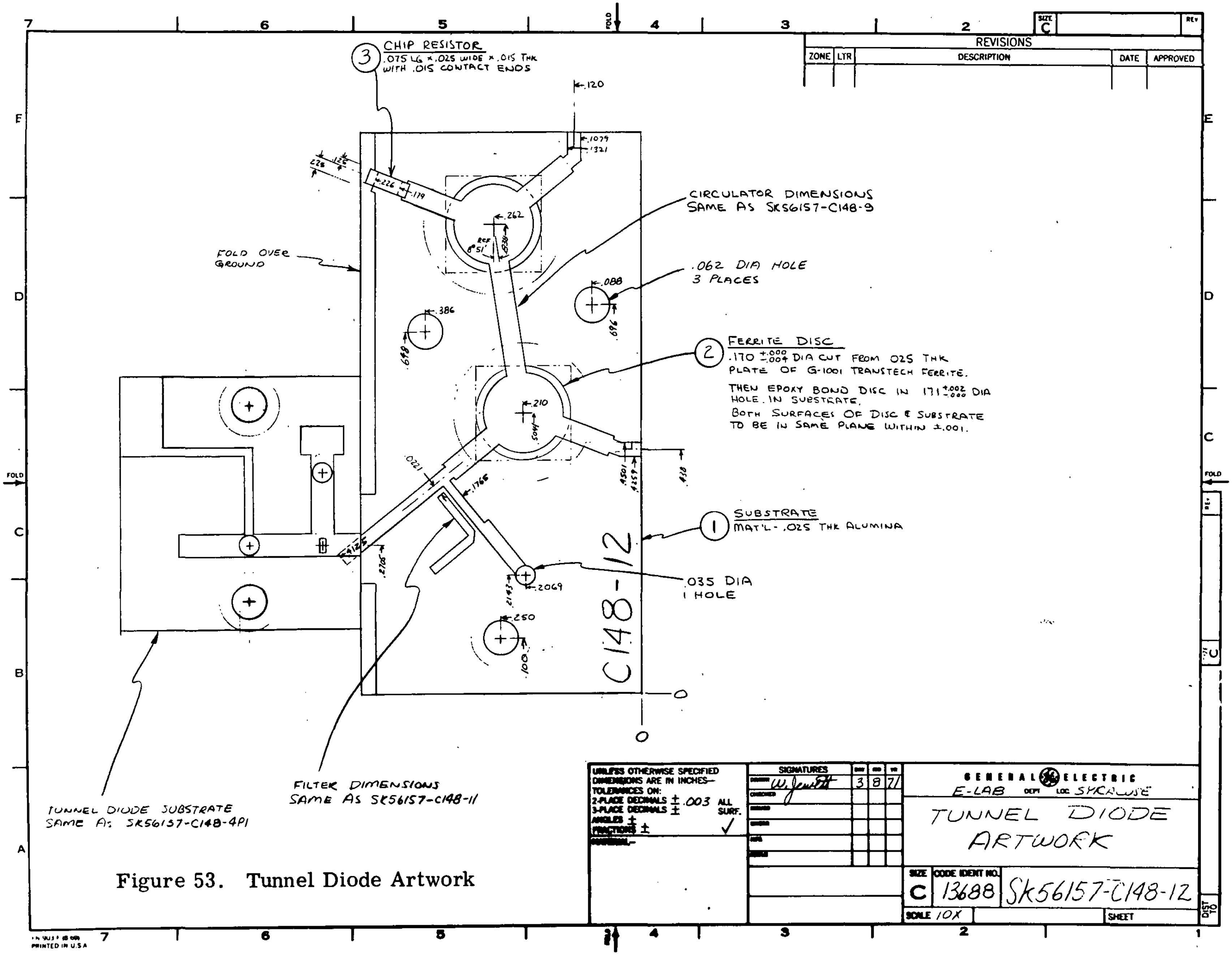


Figure 53. Tunnel Diode Artwork

TABLE XXVIII

FORWARD CHARACTERISTICS OF CIRCULATOR C-148-12
FROM PORT 1 TO PORT 2 (TD)

FREQ	REFL	ANGLE	RTN LS	VSWR	GAIN	PHASE	DELAY
11200.000	.205	179.4	13.8	1.515	-1.73	129.2	.155
11225.000	.200	178.4	14.0	1.499	-1.71	127.8	.183
11250.000	.195	177.4	14.2	1.484	-1.68	126.1	.190
11275.000	.190	176.3	14.4	1.469	-1.72	124.4	.164
11300.000	.184	175.1	14.7	1.452	-1.72	122.9	.204
11325.000	.177	174.0	15.0	1.432	-1.72	121.1	.148
11350.000	.171	173.4	15.4	1.412	-1.76	119.8	.161
11375.000	.165	172.9	15.6	1.395	-1.73	118.3	.207
11400.000	.159	172.5	16.0	1.378	-1.78	116.4	.196
11425.000	.153	171.8	16.3	1.360	-1.85	114.7	.149
11450.000	.146	171.3	16.7	1.341	-1.85	113.3	.196
11475.000	.139	171.0	17.1	1.323	-1.85	111.6	.186
11500.000	.132	171.2	17.6	1.304	-1.92	109.9	.173
11525.000	.125	171.7	18.1	1.285	-1.93	108.3	.232
11550.000	.117	172.4	18.6	1.265	-2.00	106.3	.166
11575.000	.109	174.3	19.3	1.244	-2.08	104.8	.182
11600.000	.102	177.5	19.9	1.226	-2.07	103.1	.210
11625.000	.098	-178.0	20.2	1.216	-2.15	101.2	.146
11650.000	.096	-173.5	20.4	1.212	-2.21	99.9	.189
11675.000	.096	-169.8	20.3	1.213	-2.21	98.2	.185
11700.000	.097	-166.7	20.2	1.216	-2.33	96.6	.101
11725.000	.099	-164.2	20.0	1.221	-2.38	95.6	.167
11750.000	.102	-162.9	19.9	1.226	-2.36	94.1	.123
11775.000	.103	-162.4	19.8	1.229	-2.40	93.0	.112
11800.000	.103	-162.1	19.7	1.230	-2.30	92.0	.191
11825.000	.101	-161.8	19.9	1.225	-2.24	90.4	.149
11850.000	.098	-160.8	20.2	1.217	-2.19	89.1	.177
11875.000	.094	-158.9	20.6	1.207	-2.04	87.5	.229
11900.000	.091	-155.4	20.8	1.200	-1.97	85.4	.175
11925.000	.089	-151.3	21.1	1.194	-1.90	83.8	.235
11950.000	.088	-146.3	21.1	1.194	-1.73	81.7	.279
11975.000	.090	-141.2	20.9	1.197	-1.73	79.2	.181
12000.000	.094	-136.8	20.5	1.208	-1.71	77.6	.236
12025.000	.099	-133.7	20.1	1.220	-1.65	75.4	.200
12050.000	.105	-131.1	19.6	1.234	-1.69	73.6	.198
12075.000	.111	-129.2	19.1	1.251	-1.63	71.9	.262
12100.000	.118	-127.9	18.6	1.267	-1.61	69.5	.215
12125.000	.125	-126.9	18.1	1.284	-1.67	67.6	.139
12150.000	.132	-126.4	17.6	1.303	-1.67	66.3	.246
12175.000	.138	-126.3	17.2	1.320	-1.64	64.1	.200
12200.000	.145	-126.3	16.8	1.340	-1.74	62.3	.196
12225.000	.152	-126.6	16.4	1.358	-1.70	60.5	.227
12250.000	.158	-127.0	16.0	1.376	-1.81	58.5	.159
12275.000	.165	-127.9	15.6	1.396	-1.88	57.1	.204
12300.000	.171	-128.6	15.4	1.412	-1.87	55.2	.196
12325.000	.176	-129.4	15.1	1.427	-1.98	53.5	.124
12350.000	.182	-130.4	14.8	1.445	-2.03	52.4	.247
12374.996	.188	-131.2	14.5	1.463	-2.01	50.1	.247

TABLE XXIX

FORWARD CHARACTERISTICS OF CIRCULATOR C-148-12 FROM PORT 2 (TD) TO PORT 1 (INPUT)

FREQ	REFL	ANGLE	RTN LS	VSWR	GAIN	PHASE	DELAY
11200.000	.334	-141.5	9.5	2.002	-15.69	64.3	.335
11225.000	.331	-143.0	9.6	1.991	-15.38	61.3	.455
11250.000	.329	-144.5	9.6	1.982	-16.09	57.2	.372
11275.000	.326	-146.1	9.7	1.969	-16.43	53.7	.356
11300.000	.323	-147.6	9.8	1.953	-16.65	50.5	.459
11325.000	.319	-149.2	9.9	1.937	-16.95	46.4	.369
11350.000	.314	-150.4	10.1	1.915	-17.29	43.1	.397
11375.000	.312	-151.7	10.1	1.903	-17.50	39.5	.454
11400.000	.309	-153.1	10.2	1.894	-17.85	35.4	.409
11425.000	.304	-154.5	10.3	1.874	-18.20	31.7	.413
11450.000	.302	-155.7	10.4	1.864	-18.44	28.0	.472
11475.000	.298	-156.9	10.5	1.848	-18.80	23.7	.396
11500.000	.295	-158.1	10.6	1.836	-19.19	20.2	.525
11525.000	.293	-159.0	10.7	1.823	-19.49	15.4	.531
11550.000	.290	-160.1	10.8	1.816	-20.02	10.7	.500
11575.000	.287	-161.0	10.8	1.806	-20.56	6.2	.507
11600.000	.285	-162.0	10.9	1.797	-21.17	1.6	.426
11625.000	.284	-162.8	10.9	1.794	-21.83	-2.2	.403
11650.000	.283	-163.6	11.0	1.791	-22.39	-5.9	.502
11675.000	.284	-164.6	10.9	1.792	-22.99	-10.4	.523
11700.000	.283	-165.6	11.0	1.791	-23.64	-15.1	.519
11725.000	.284	-166.7	10.9	1.795	-24.22	-19.8	.561
11750.000	.285	-168.0	10.9	1.796	-24.86	-24.8	.567
11775.000	.284	-169.6	10.9	1.791	-25.43	-29.9	.587
11800.000	.284	-170.9	10.9	1.793	-25.99	-35.2	.732
11825.000	.280	-172.3	11.0	1.779	-26.43	-41.8	.735
11850.000	.277	-173.4	11.1	1.767	-26.77	-48.4	.862
11875.000	.274	-174.6	11.2	1.755	-26.97	-56.2	.976
11900.000	.270	-175.4	11.4	1.741	-27.20	-64.9	.897
11925.000	.267	-176.0	11.5	1.729	-27.20	-73.0	1.107
11950.000	.264	-176.6	11.6	1.718	-26.94	-83.0	1.146
11975.000	.261	-177.0	11.7	1.707	-26.83	-93.3	1.036
12000.000	.260	-177.2	11.7	1.703	-26.56	-102.6	1.029
12025.000	.259	-177.6	11.8	1.697	-26.14	-111.9	.935
12050.000	.258	-177.9	11.8	1.695	-25.72	-120.3	.904
12075.000	.258	-178.4	11.8	1.697	-25.19	-128.4	.802
12100.000	.257	-178.3	11.8	1.693	-24.72	-135.4	.726
12125.000	.257	-179.3	11.8	1.693	-24.17	-143.0	.695
12150.000	.258	-179.6	11.8	1.697	-23.66	-149.2	.673
12175.000	.259	179.6	11.7	1.698	-23.11	-155.3	.670
12200.000	.260	178.9	11.7	1.702	-22.59	-161.3	.614
12225.000	.259	178.2	11.7	1.701	-22.04	-166.8	.644
12250.000	.260	177.5	11.7	1.703	-21.61	-172.6	.571
12275.000	.260	176.7	11.7	1.703	-21.14	-177.8	.556
12300.000	.260	175.8	11.7	1.704	-20.68	177.2	.565
12325.000	.260	174.8	11.7	1.702	-20.28	172.1	.516
12350.000	.260	173.7	11.7	1.704	-19.85	167.5	.559
12374.996	.260	172.3	11.7	1.702	-19.43	162.5	.559

TABLE XXX

FORWARD CHARACTERISTICS OF CIRCULATOR C-148-12 FROM PORT 4 (OUTPUT) TO PORT 2 (TD)

FREQ	REFL	ANGLE	RTN LS	VSWR	GAIN	PHASE	DELAY
11200.000	.226	130.7	12.9	1.583	-34.78	-34.2	.456
11225.000	.222	129.4	13.1	1.570	-35.15	-38.3	.559
11250.000	.217	128.3	13.3	1.556	-35.56	-43.3	.573
11275.000	.214	127.1	13.4	1.543	-36.14	-48.5	.491
11300.000	.210	126.2	13.6	1.530	-36.77	-52.9	.587
11325.000	.206	125.2	13.7	1.518	-37.32	-58.2	.599
11350.000	.202	124.5	13.9	1.505	-37.83	-63.6	.692
11375.000	.197	123.7	14.1	1.491	-38.57	-69.8	.804
11400.000	.192	123.2	14.3	1.474	-39.43	-77.1	.809
11425.000	.186	122.6	14.6	1.457	-40.14	-84.3	1.107
11450.000	.181	122.1	14.8	1.443	-41.20	-94.3	1.371
11475.000	.176	121.9	15.1	1.423	-42.27	-106.6	1.722
11500.000	.171	122.1	15.3	1.412	-43.80	-122.1	2.039
11525.000	.167	122.6	15.5	1.401	-45.66	-140.5	3.794
11550.000	.164	123.5	15.7	1.392	-48.72	-174.6	5.299
11575.000	.163	124.0	15.7	1.391	-52.41	137.7	6.828
11600.000	.163	123.2	15.8	1.388	-53.39	76.3	4.307
11625.000	.160	122.3	15.9	1.381	-52.18	37.5	2.645
11650.000	.155	121.6	16.2	1.368	-50.61	13.7	.981
11675.000	.151	121.4	16.4	1.356	-50.85	4.9	1.154
11700.000	.146	121.3	16.7	1.342	-50.85	-5.5	.065
11725.000	.141	121.5	17.0	1.329	-50.73	-4.9	.183
11750.000	.137	122.2	17.3	1.318	-50.37	-6.6	.088
11775.000	.133	122.4	17.5	1.306	-49.80	-5.8	.623
11800.000	.129	122.7	17.8	1.295	-49.34	-.2	.074
11825.000	.125	123.7	18.1	1.286	-48.17	-.8	.050
11850.000	.122	124.6	18.3	1.277	-46.61	-1.3	.399
11875.000	.118	125.1	18.6	1.268	-44.98	-4.9	.596
11900.000	.114	126.2	18.8	1.258	-44.14	-10.2	.519
11925.000	.111	127.4	19.1	1.249	-42.82	-14.9	.834
11950.000	.108	129.1	19.4	1.241	-41.65	-22.4	.945
11975.000	.104	131.0	19.7	1.232	-40.69	-30.9	.897
12000.000	.101	133.4	19.9	1.225	-40.07	-39.0	1.030
12025.000	.099	136.1	20.1	1.219	-39.52	-48.3	1.064
12050.000	.097	139.1	20.3	1.214	-38.88	-57.9	.994
12075.000	.096	142.7	20.3	1.213	-38.37	-66.8	.967
12100.000	.096	145.7	20.3	1.214	-37.96	-75.5	1.026
12125.000	.097	148.8	20.2	1.215	-37.56	-84.7	.779
12150.000	.099	152.0	20.1	1.220	-37.15	-91.7	1.083
12175.000	.102	154.5	19.8	1.227	-36.61	-101.5	.993
12200.000	.105	156.8	19.5	1.236	-36.32	-110.4	.888
12225.000	.109	158.8	19.2	1.246	-35.73	-118.4	.924
12250.000	.114	160.3	18.9	1.257	-35.44	-126.7	.794
12275.000	.118	161.4	18.5	1.269	-34.96	-133.9	.860
12300.000	.124	162.2	18.2	1.282	-34.36	-141.6	.970
12325.000	.129	162.5	17.8	1.297	-34.16	-150.4	.898
12350.000	.135	162.5	17.4	1.311	-33.85	-158.4	.959
12374.996	.140	162.0	17.1	1.326	-33.54	-167.1	.959

TABLE XXXI

FORWARD CHARACTERISTICS OF CIRCULATOR C-148-12 FROM PORT 1 (INPUT) TO PORT 4 (OUTPUT) WITH PORT 2 OPEN CIRCUITED

FREQ	REFL	ANGLE	RTN LS	VSWR	GAIN	PHASE	DELAY
11200.000	.323	-155.2	9.7	1.975	-3.73	-44.5	.330
11225.000	.324	-158.1	9.8	1.957	-3.69	-47.5	.441
11250.000	.317	-161.1	10.0	1.928	-3.64	-51.5	.394
11275.000	.308	-164.0	10.2	1.891	-3.74	-54.9	.393
11300.000	.299	-166.6	10.5	1.854	-3.72	-58.5	.444
11325.000	.290	-169.1	10.7	1.818	-3.76	-62.5	.304
11350.000	.281	-171.2	11.0	1.780	-3.75	-65.2	.447
11375.000	.275	-173.8	11.2	1.757	-3.70	-69.2	.455
11400.000	.265	-176.1	11.5	1.722	-3.83	-73.3	.375
11425.000	.256	-178.4	11.8	1.688	-3.90	-76.8	.405
11450.000	.247	179.3	12.2	1.655	-3.89	-80.4	.412
11475.000	.237	177.2	12.5	1.622	-3.97	-84.1	.401
11500.000	.228	175.0	12.8	1.590	-4.01	-87.8	.439
11525.000	.217	173.2	13.3	1.555	-4.04	-91.7	.410
11550.000	.206	171.5	13.7	1.519	-4.14	-95.4	.383
11575.000	.196	170.4	14.2	1.486	-4.14	-98.8	.437
11600.000	.184	169.2	14.7	1.452	-4.18	-102.8	.406
11625.000	.175	168.3	15.2	1.423	-4.28	-106.4	.329
11650.000	.165	168.0	15.6	1.396	-4.31	-109.4	.381
11675.000	.156	168.0	16.2	1.369	-4.30	-112.8	.337
11700.000	.148	168.3	16.6	1.346	-4.37	-115.9	.302
11725.000	.141	168.7	17.0	1.329	-4.32	-118.6	.319
11750.000	.136	168.6	17.4	1.313	-4.23	-121.5	.295
11775.000	.130	168.2	17.7	1.298	-4.09	-124.1	.359
11800.000	.124	167.5	18.1	1.284	-3.84	-127.3	.405
11825.000	.118	166.5	18.6	1.267	-3.67	-131.0	.414
11850.000	.110	165.2	19.2	1.247	-3.43	-134.7	.489
11875.000	.100	163.6	20.0	1.221	-3.21	-139.1	.521
11900.000	.089	163.0	21.0	1.195	-3.09	-143.8	.502
11925.000	.076	163.4	22.4	1.164	-3.00	-148.3	.523
11950.000	.063	166.4	24.0	1.135	-2.92	-153.0	.523
11975.000	.051	172.8	25.8	1.103	-2.87	-157.7	.515
12000.000	.042	-176.0	27.6	1.087	-2.89	-162.4	.560
12025.000	.037	-157.3	28.7	1.076	-2.92	-167.4	.494
12050.000	.037	-136.5	28.6	1.077	-3.02	-171.9	.529
12075.000	.045	-119.6	27.0	1.094	-3.07	-176.6	.5 2
12100.000	.056	-110.3	25.0	1.119	-3.16	178.1	.547
12125.000	.069	-105.6	23.2	1.149	-3.33	173.1	.503
12150.000	.083	-103.5	21.6	1.181	-3.46	168.6	.528
12175.000	.097	-102.9	20.3	1.214	-3.55	163.9	.552
12200.000	.111	-103.3	19.1	1.249	-3.69	158.9	.559
12225.000	.124	-105.0	18.1	1.283	-3.82	153.9	.511
12250.000	.137	-106.3	17.2	1.319	-4.01	149.3	.481
12275.000	.150	-108.8	16.5	1.353	-4.15	144.9	.528
12300.000	.163	-110.8	15.7	1.390	-4.22	140.2	.547
12325.000	.175	-113.2	15.1	1.424	-4.39	135.3	.435
12350.000	.188	-115.6	14.5	1.463	-4.57	131.3	.500
12374.996	.199	-118.0	14.0	1.497	-4.63	126.9	.502

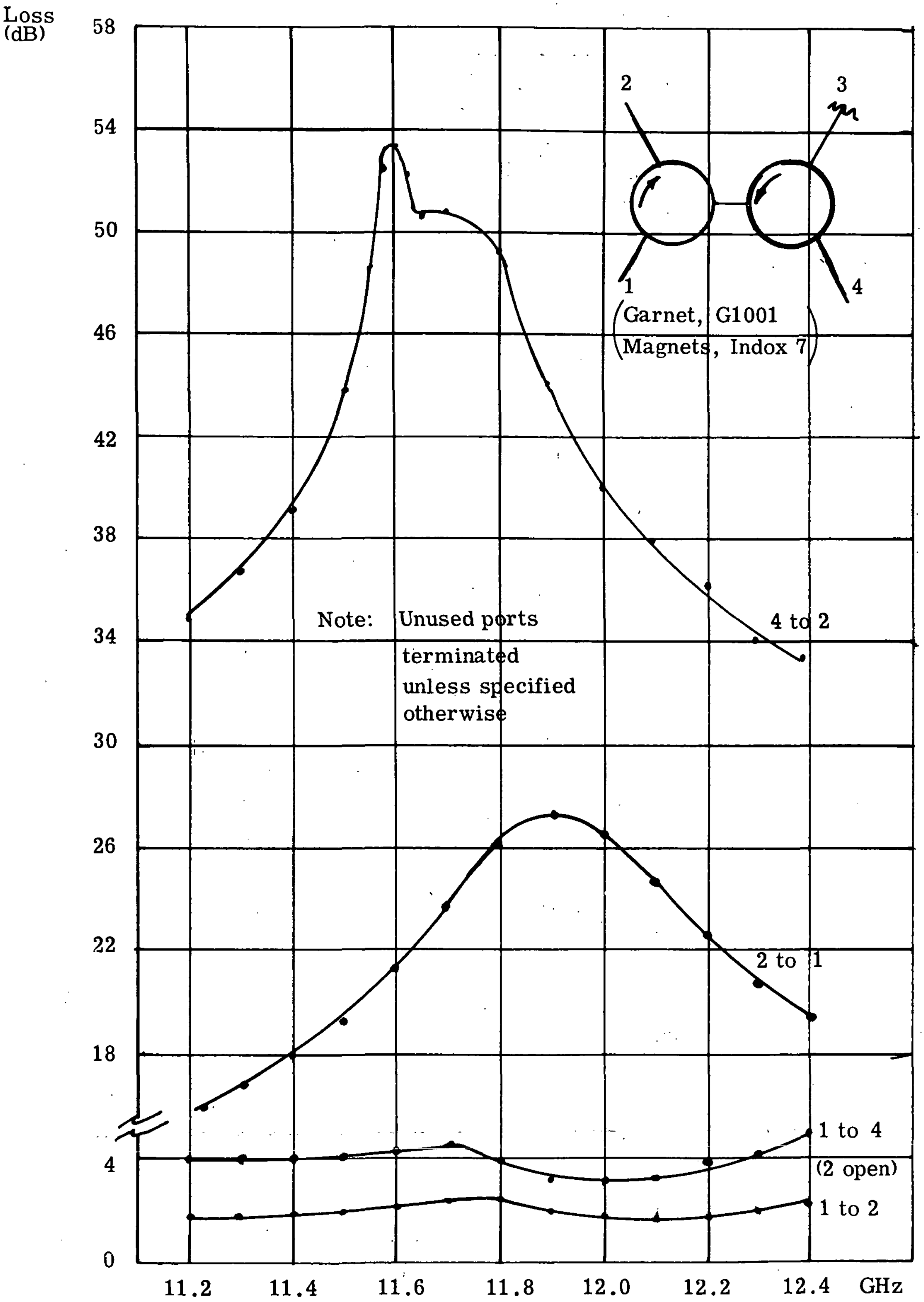


Figure 54. Characteristics of Circulator C-148-12

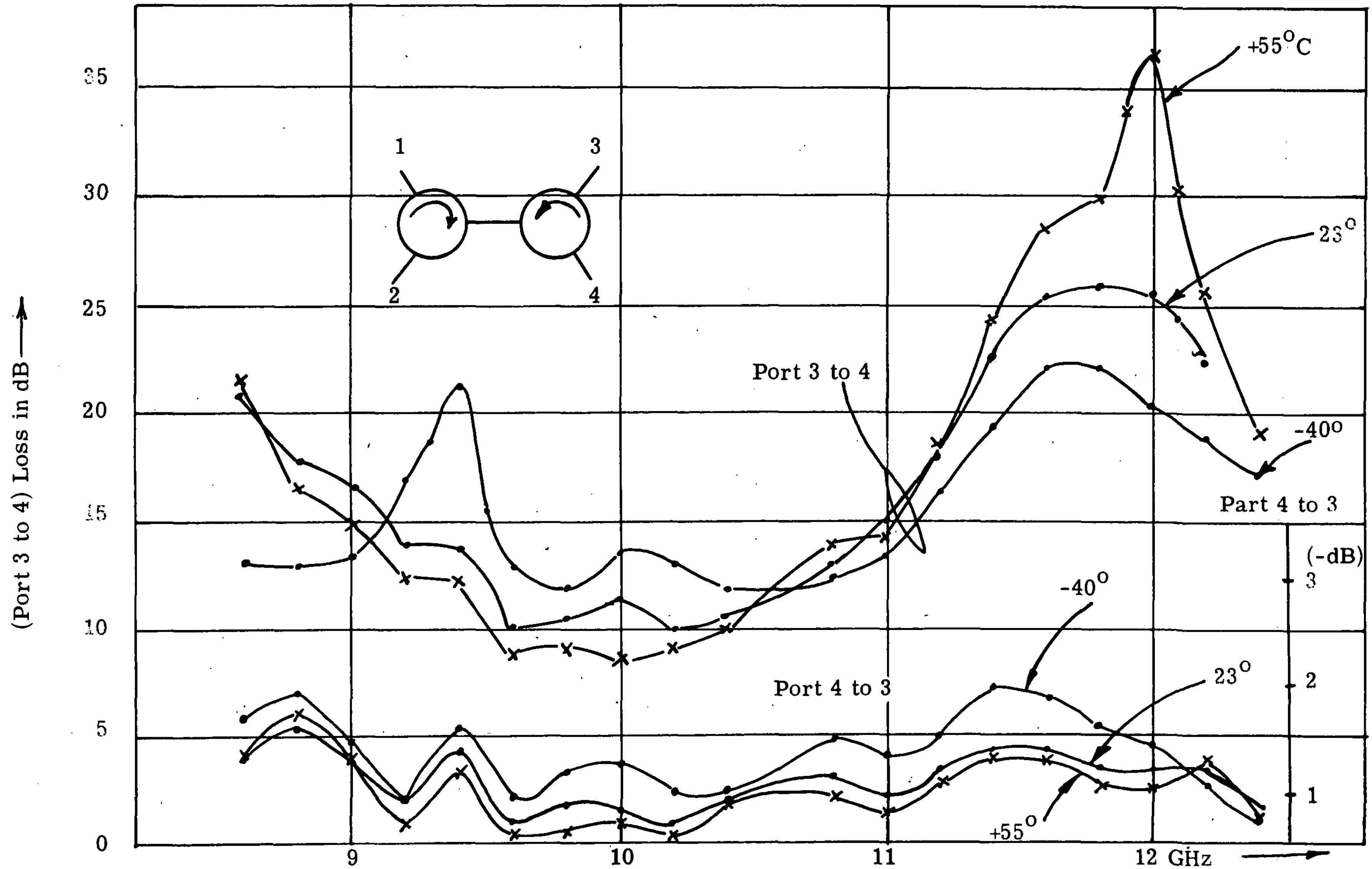


Figure 55. Thermal Effect on Circulator Characteristics Unit Design C-148-9

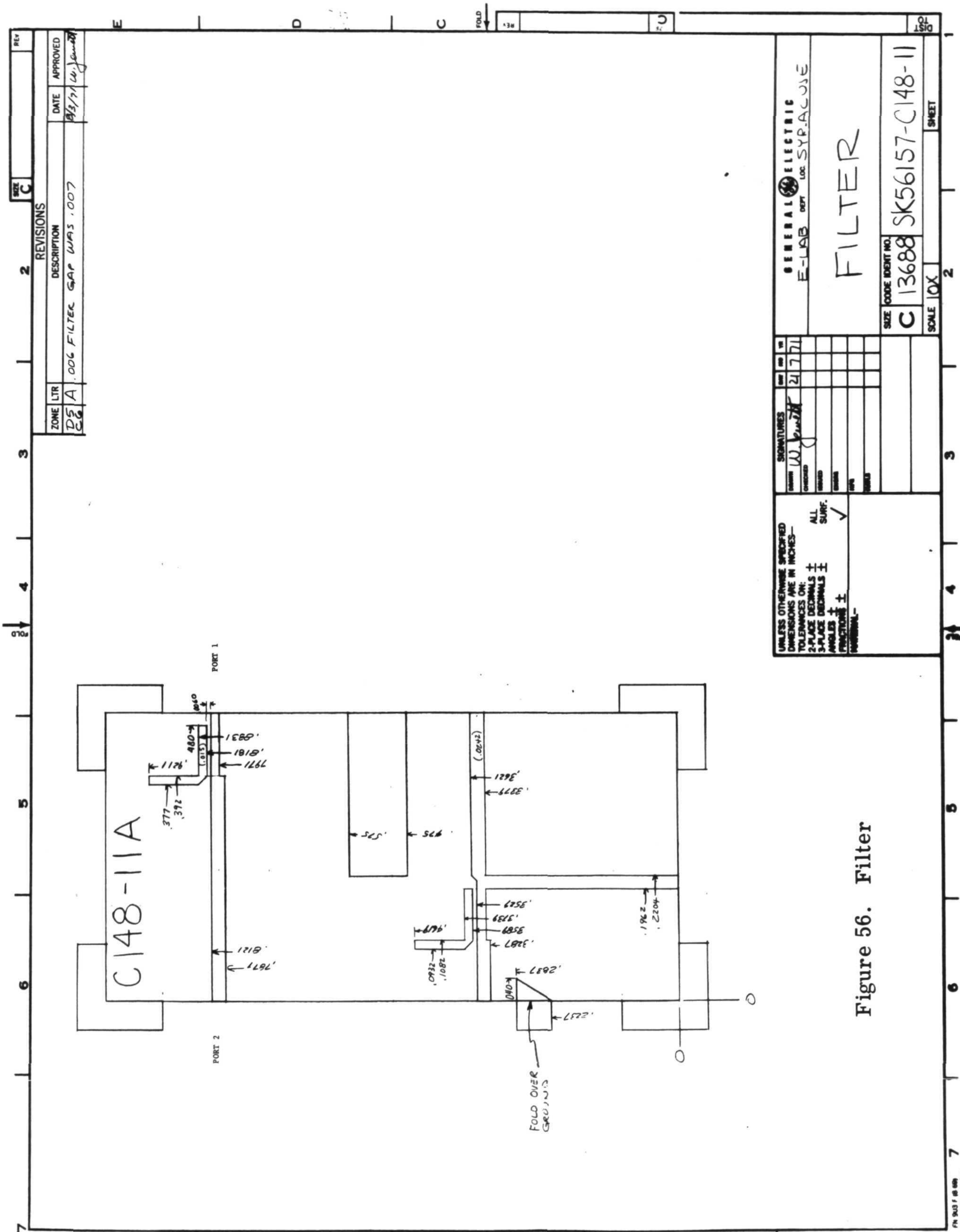
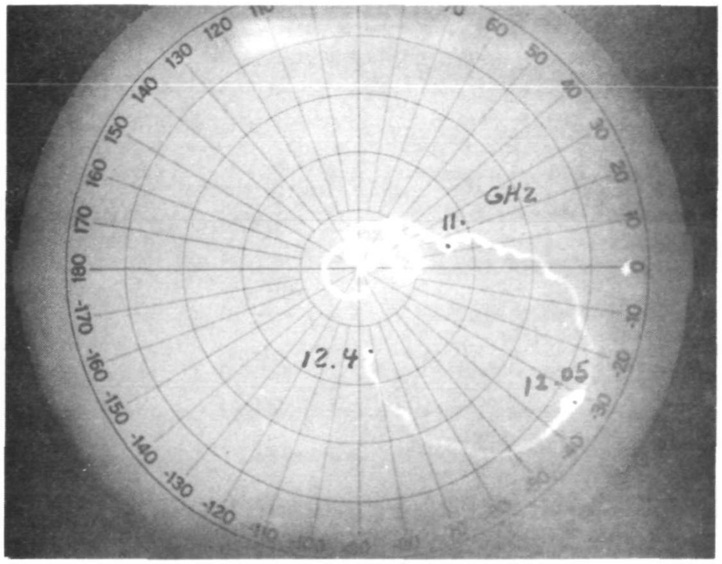
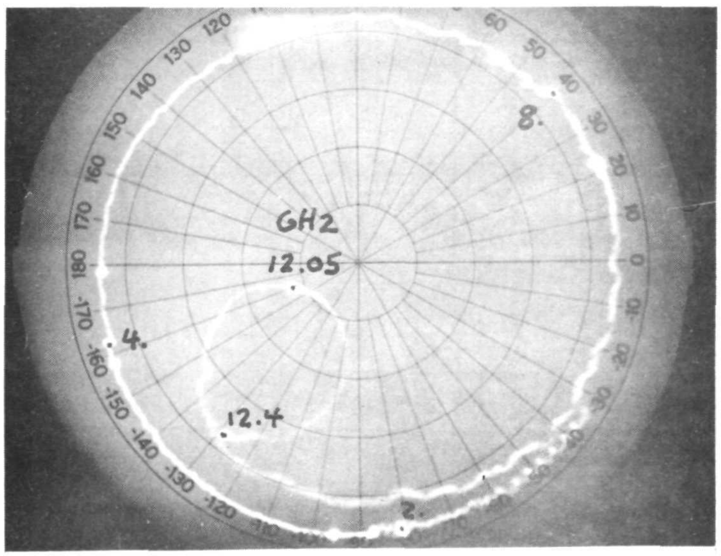


Figure 56. Filter



(a) Reflection Coefficient



(b) Transmission Characteristics

Figure 57. Filter Characteristics

6.6 T.D. SUBSTRATE

6.6.1 Design

The layout of the T.D. substrate is shown in Figure 58. The stabilizing circuit, as shown, provides two shunt branches across the tunnel diode. The branch to the right of the T.D. has two $\lambda/4$ sections in tandem and one 50 ohm resistor to ground at the junction where the dc supply line is connected. (Since the lines are dc grounded in the overall setup, this feeder line can be ignored). This branch presents an open or high impedance to the T.D. at the center of the passband and a resistive admittance at low frequencies. The second line serves a similar function in the passband and the low frequency band, but will stagger the possible resonant frequencies of the two branches at higher harmonics, for example, near the second harmonic.

The admittance of the stabilizing circuit, including two 50-ohm silicon chip resistors, measured at the site of the T.D., is shown in Figure 59. The expected characteristic should be a smooth curve without the minor loops, A and B, as shown. These loops are probably due to the presence of the dc supply feeder in combination with the branches. The criteria of the stabilization circuit are that the total conductance, including the load across the terminals of the T.D., should remain positive; or the total susceptance does not pass through zero even though the conductance becomes negative. As the admittance plot shows, it is fairly safe to say that, below 7 GHz, there is no danger of oscillation even if the load is reactive and away from the nominal load of 50 ohms, because the negative conductance of the T.D. is in the region of $0.9 \times 1/50$ mhos (absolute magnitude). But, anywhere between 7 and 11.0 GHz, conditions may permit parasitic oscillation if the load happens to have certain reactive values. The filter discussed in the previous section would be effective in improving stability if located at the T.D. terminals. But at some distance away the situation is not clear.

6.6.2 Prototype Fabrication

The T.D. substrate fabrication procedure is as follows:

(1) Fuse K650 glass and the germanium bar, locating the germanium pellet within the tolerance for the final layout.

(2) Slice the above in 0.025" thick slabs.

(3) Size the slices, with reference to location and the orientation of the Ge pellet.

(4) Gold alloy the back of the Ge pellet on the side of the ground plane to provide an ohmic contact.

(5) Metallize the top and bottom of the above, with chrome, gold films and, then gold plating.

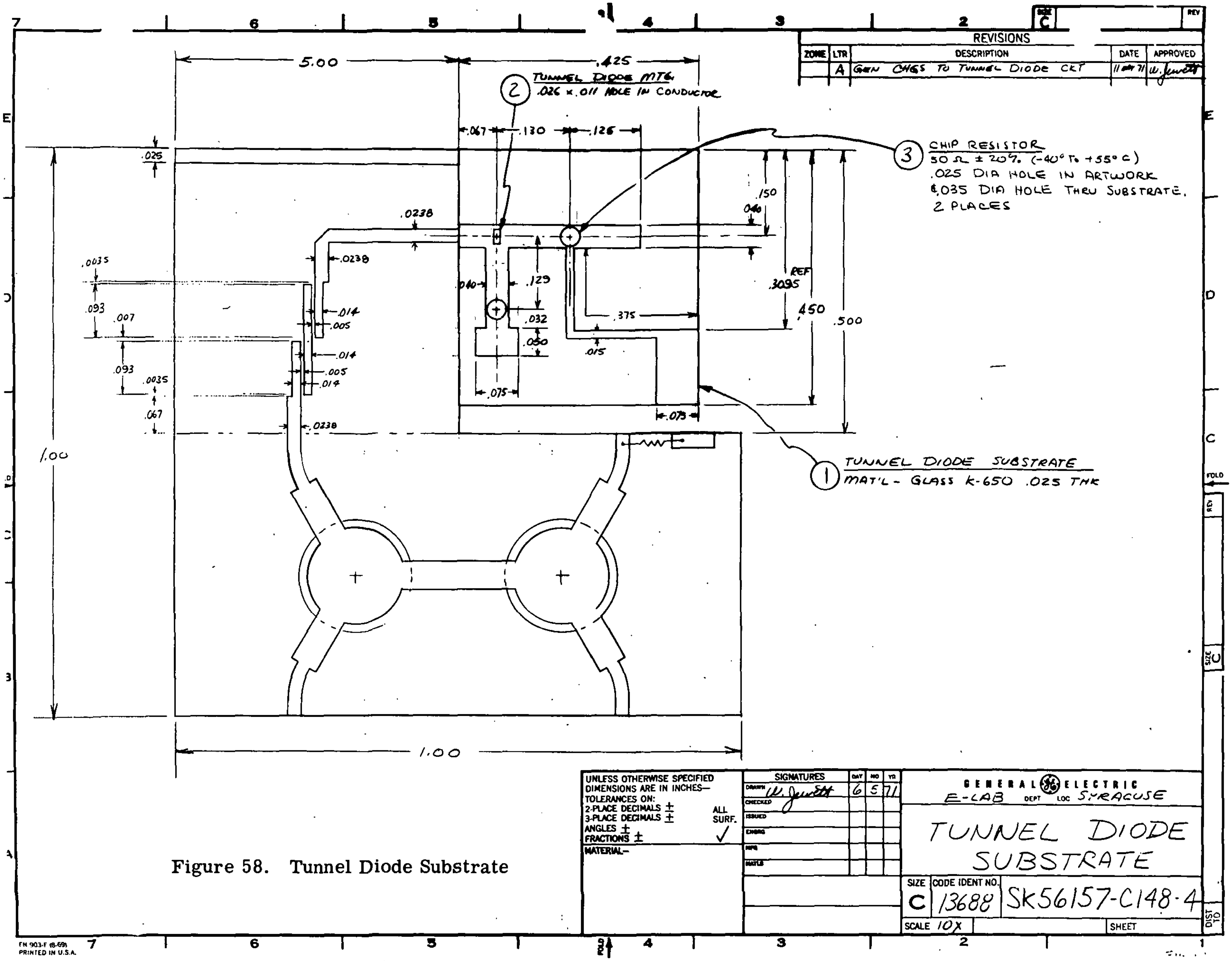


Figure 58. Tunnel Diode Substrate

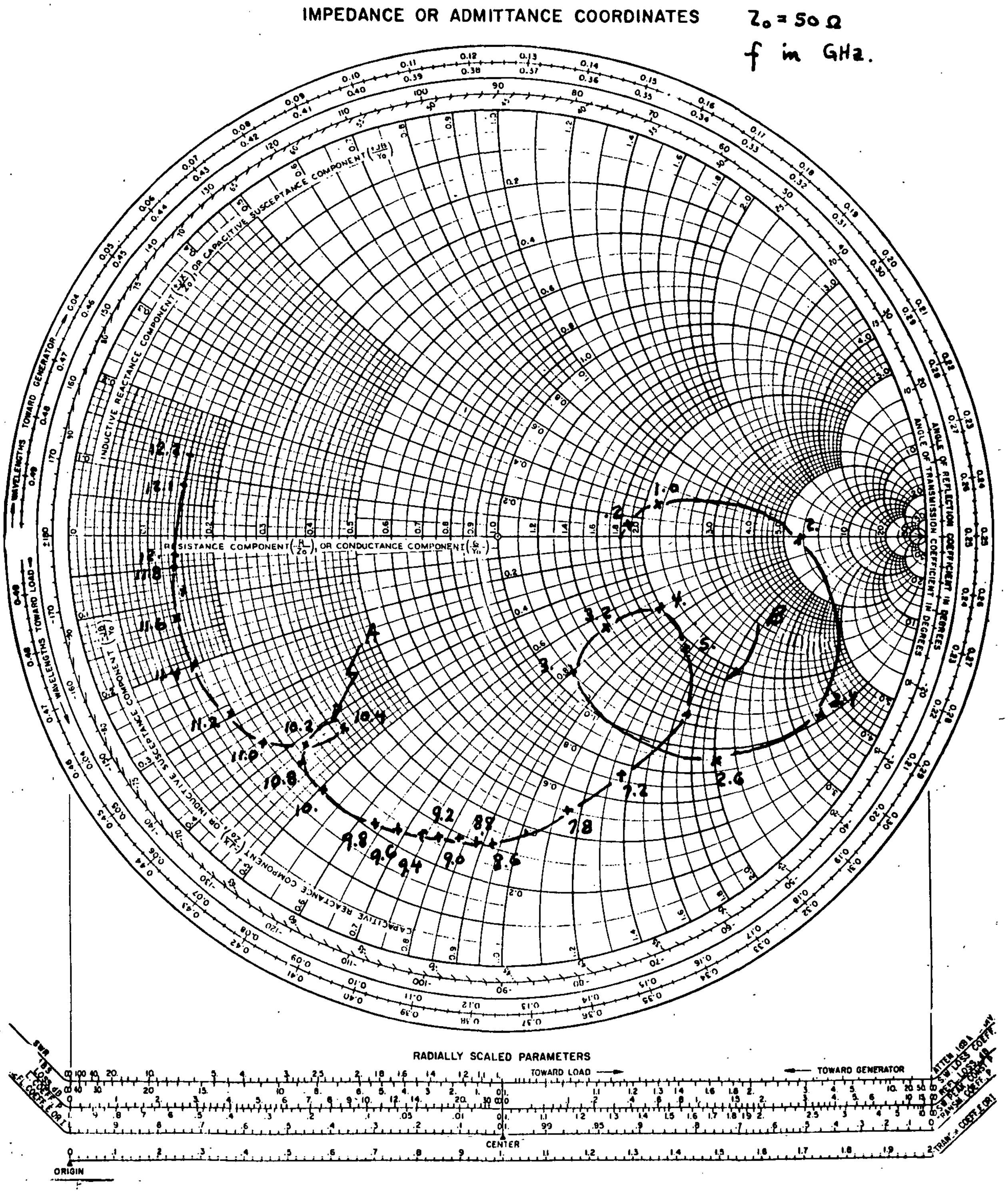


Figure 59. Admittance - Stabilizing Circuit

(6) Etch the pattern by photomask techniques.

(7) Drill the holes for the 50-ohm silicon pill resistors.

(8) Form the diode junction.

(9) Evaluate the diode characteristics

(10) Attach the resistors by wire bonding or soldering or both.

The most difficult step of the above is of course item (8), forming the low capacitance tunnel junctions, requiring a difficult alloying and etching procedure. An idea of the difficulty is reflected in the current price of these microwave T. D. 's; they cost from $270 to $300 per device.

The following data were recorded on the six units out of the first batch of 10.

Unit No.	I_p (ma.)	I_v (ma.)	v_p (mv.)	v_v (mv.)	v_F (mv.)	C_v (pf)
1-2	1.96	0.17	79	410	580	0.3
1-3	1.66	0.19	106	440	630	0.1
1-4	2.1	2.1	82	425	580	0.3
2-3	2.1	2.1	88	440	615	0,15
2-4	1.92	1.92	78	420	580	0.25
2-6	2.1	2.1	82	440	606	0.20

Because of the integrated structure, the junction capacitances were measured by taking the difference between the total capacitances before and after fabrication. The accuracy is about ±0.05 pf.

Later batches of small numbers ran into more trouble and very low yields.

The spread of the junction capacitance poses a serious problem, for the TDA design, in finding proper matching sections. It is likely that no fixed matching sections will do the job for the 1 to 3 capacitance spread. Even with selection, one still needs some variable tuning. In the current design the dc bias current is used to perform minor tuning.

6.7 DC BIAS

The dc bias design is quite trivial. The schematic is shown on Figure 60. Note that the TD substrate ground plane is insulated from the package by a piece of 2 mil thick mylar so that a positive voltage supply and a p-type Ge pellet can be used. The total dc bias current through the diode and the stabilization resistors is about 5.6 ma. With the Zener diode used, the total dc current for the initial setting may be done by a potentimeter. It should also be noted

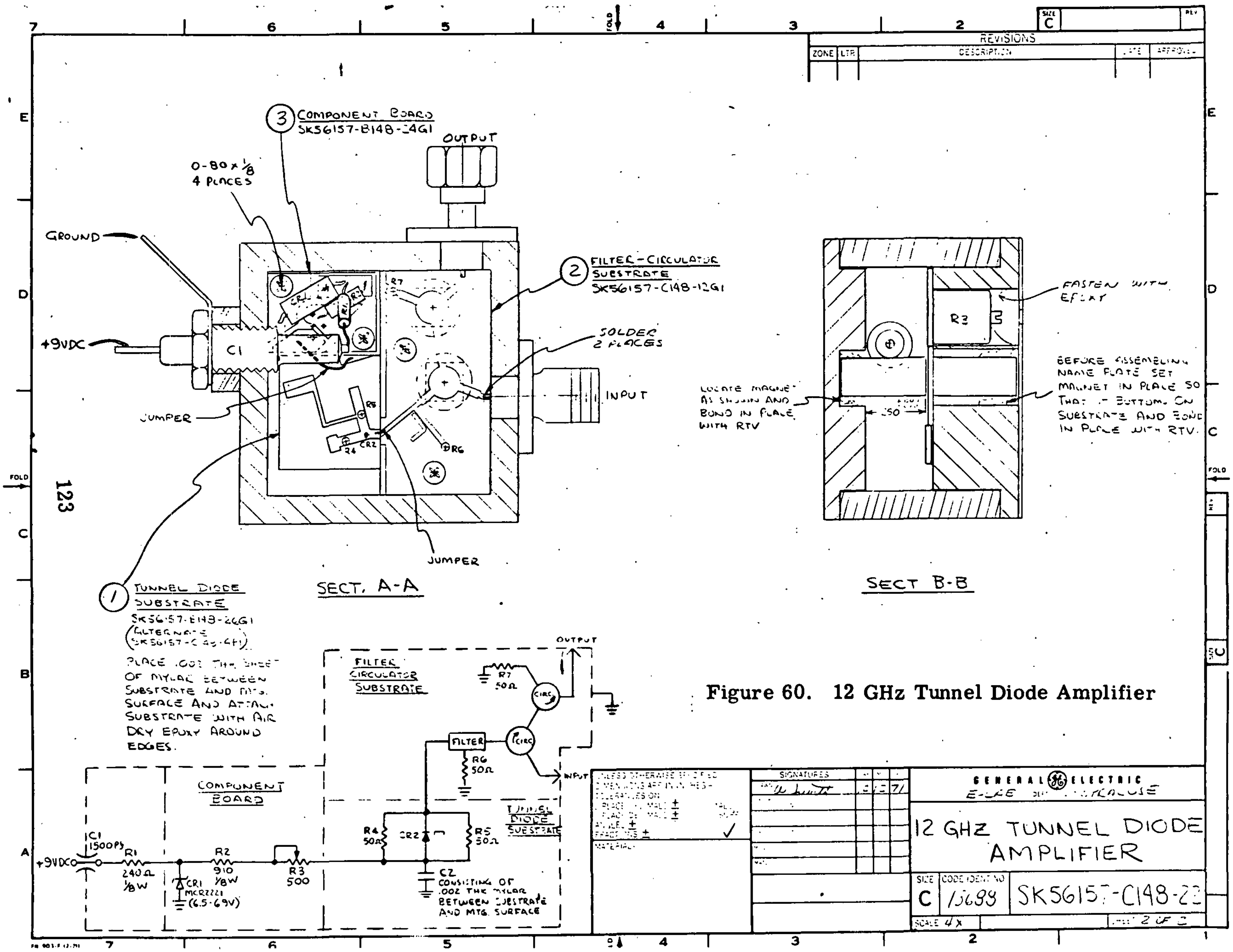


Figure 60. 12 GHz Tunnel Diode Amplifier

that the bias current through the diode depends on the dc load of the output port, whether a short or 50 ohms. Proper attention in use is required.

6.8 TEST RESULTS

Two of the assumbled units, with TD junction capacities of 0.15 pf, and designed per Figure 61, exhibited the best performance. The other units, with capacitance of about 0.20 pf, did not do as well.

An alternate design, (see Figure 62), where the stabilizing lines are foreshortened to offset the diode junction capacitance, oscillates strongly.

For the unit with 0.15 pf junction capacitance, the bias current could be adjusted to yield a steady amplification, although the adjustment range was small. Evidently, near the middle portion of the negative conductance region, the unit tends to burst into oscillation. The overall gain was recorded using a sweep generator and a spectrum analyzer, and is shown in Figure 63. The gain was above 10 dB and the bandwidth was over 100 MHz. It was not known, then, that the gain was signal-level dependent. When the signal was reduced to below -50 dBm there was a drop in gain, probably because of the marginal instability and the non-linear properties of the TD characteristic.

An attempt to measure the TDA characteristics from port to port using network analyzer yielded the result shown in Table XXXII. It did not agree with the spectrum analyzer results obtained previously. The spectrum analyzer gives a reading from 1 to 2 dB higher than that given by the network analyzer although near 12 GHz they are about the same. This was, unfortunately, not taken seriously until the signal level discrepancy was confirmed.

It was obvious, then, that the design was faulty. It was recognized that unless the marginal stability was eliminated, the gain-bandwidth would not be satisfactory.

Unfortunately, the hardware had proceeded to the final stage and the number of available TD substrates was limited. Not much could be done. The possibility of changing the matching condition between the circulator and TD substrates was investigated. A temporary matching, with metallized BeO chips placed over the line, was tried and it was encouraging that, indeed, the gain and bandwidth could be optimized with no instabilities. The photos, shown previously as Figures 49a. and b., display the operation with an input signal level of -53 dBm. It had a center-band gain of about 10 dB and a bandwidth of well over 400 MHz.

This unit was untimely damaged in the process of making some mechanical changes; the diode was found to be short circuited.

A continuous effort under G.E. internal funding to determine the mechanism of matching, and a scheme of using discrete devices in a strip-line configuration was considered the only avenue open. This work is described in 6.10, Addendum. A promising design procedure, using an extended Smith Chart, is proposed; a computed result looks promising. However, at the time of this writing the experimental effort has not been completed.

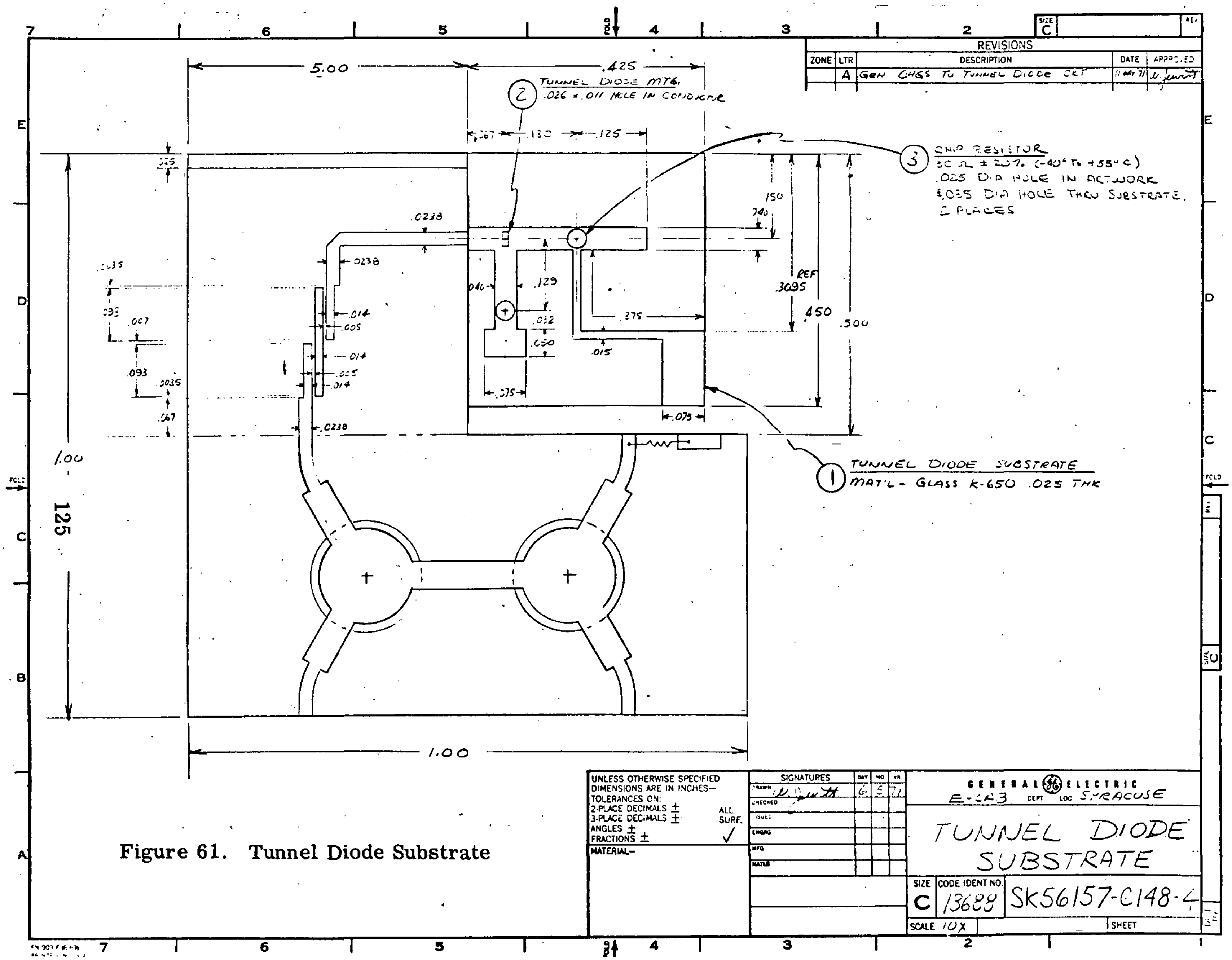


Figure 61. Tunnel Diode Substrate

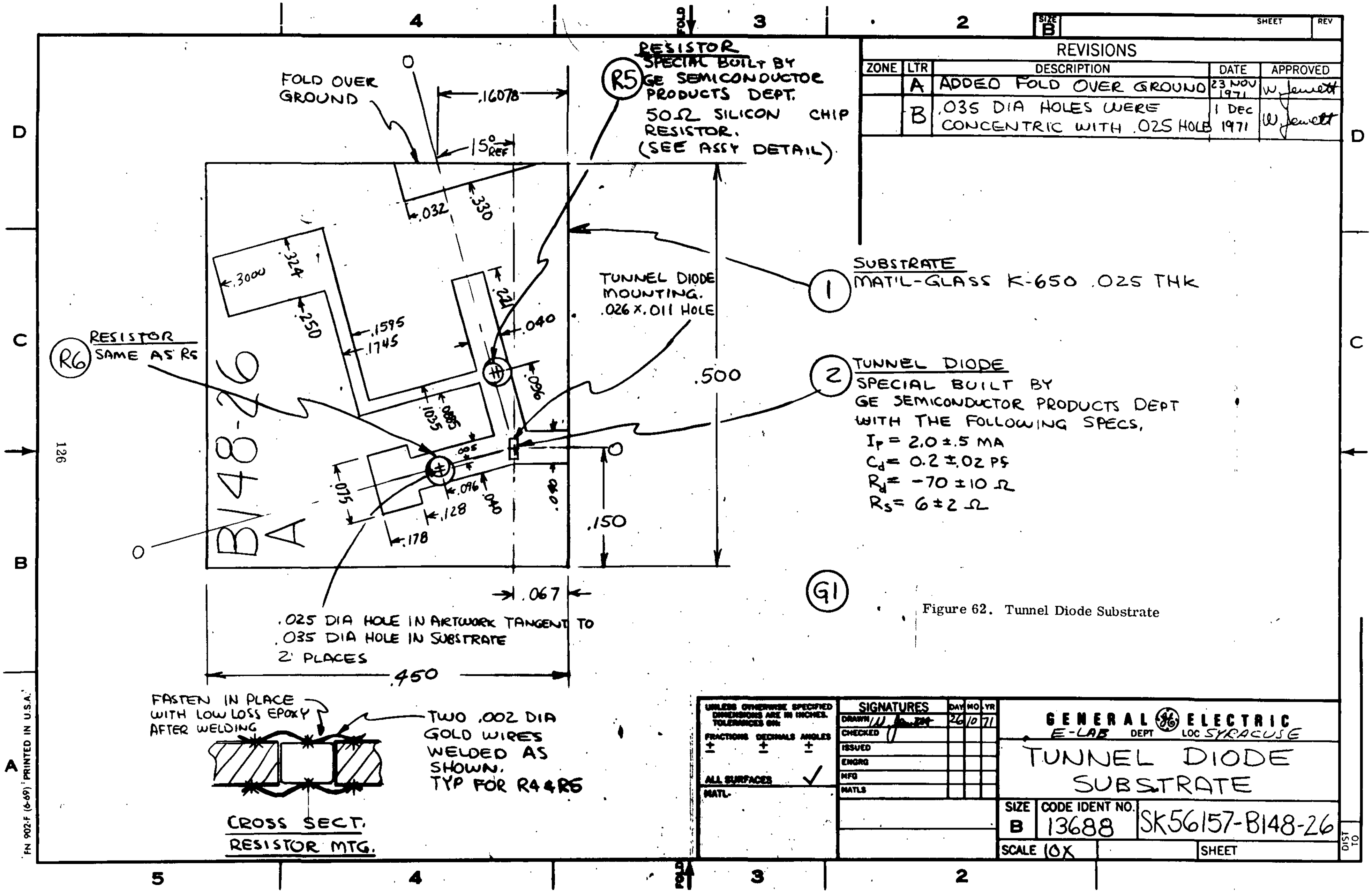
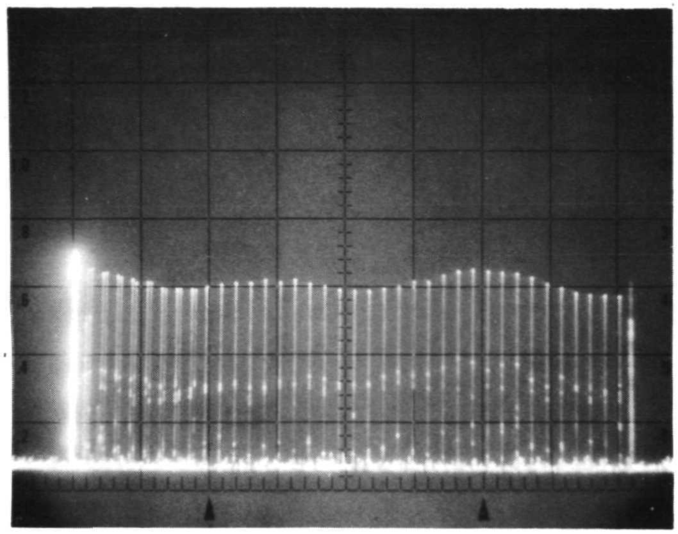


Figure 62. Tunnel Diode Substrate



Vertical 10 dB/cm; Horizontal 30 MHz/cm

Figure 63. TDA Gain Display for Unit #2

TABLE XXXII.

UNIT #1 MEASURED ON NETWORK ANALYZER

TASK? 2

CONN DEVICE -- BILATERAL? N TDA I 5.7 MA

FREQ	REFL	VSWR	ANGLE	GAIN	PHASE
11880.000	.727	6.319	26.7	11.99	46.3
11890.000	.704	5.767	19.0	11.84	39.3
11900.000	.674	5.139	12.5	11.90	32.6
11910.000	.658	4.852	3.2	11.91	24.7
11920.000	.617	4.220	-4.2	11.86	17.1
11930.000	.551	3.458	-12.1	11.98	10.1
11940.000	.509	3.071	-21.1	11.80	1.8
11950.000	.478	2.830	-30.2	11.57	-6.0
11960.000	.451	2.646	-39.3	11.19	-13.2
11970.000	.403	2.353	-52.0	10.86	-20.2
11980.000	.373	2.191	-60.8	10.45	-26.5
11990.000	.344	2.048	-71.3	10.14	-32.6
12000.000	.329	1.981	-83.7	9.86	-38.9
12010.000	.310	1.898	-94.2	9.51	-45.1
12020.000	.313	1.909	-107.0	9.13	-51.4
12030.000	.294	1.834	-111.0	8.59	-56.0
12040.000	.310	1.897	-121.8	8.22	-61.0
12050.000	.294	1.835	-127.8	7.87	-65.0
12060.000	.311	1.902	-135.5	7.56	-68.9
12070.000	.317	1.928	-139.4	7.35	-73.4
12080.000	.331	1.990	-145.2	7.10	-78.8
12090.000	.325	1.963	-149.4	6.71	-83.9
12100.000	.334	2.002	-153.9	6.13	-88.4
12110.000	.336	2.010	-155.6	5.69	-92.3
12119.998	.343	2.043	-159.2	5.19	-95.3

TASK? 2

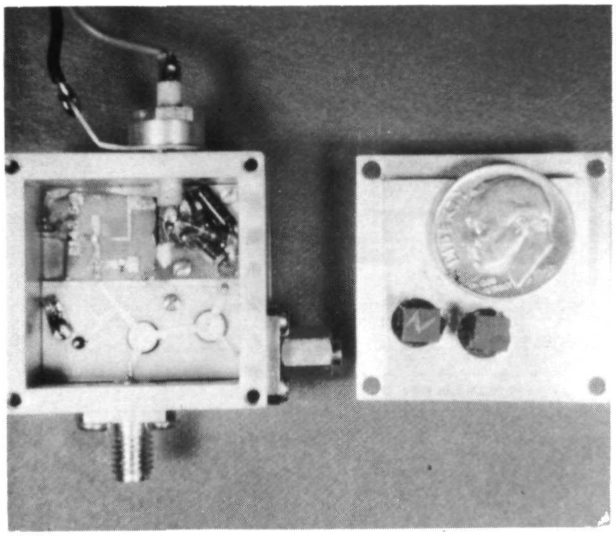
The noise figure measurement with various approaches indicates that, approximately, 6 to 7 dB can be accomplished.

6.9 PACKAGING DETAILS AND COST ESTIMATES

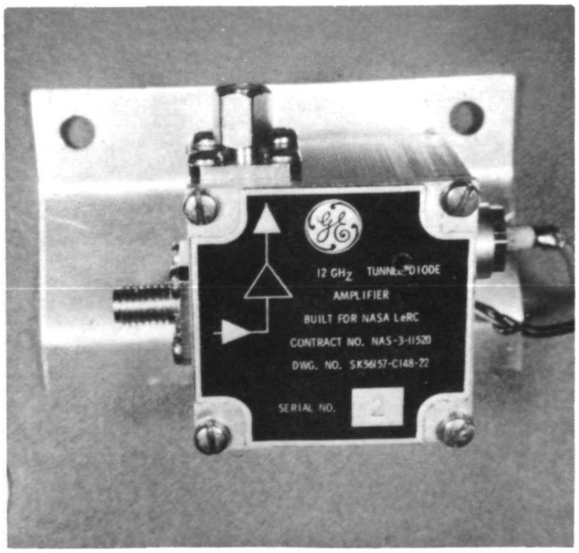
The final packaging scheme is shown in Figures 64a and b. The mechanical drawings listed below spell out complete details.

The general packaging design will remain the same even if additional matching devices are introduced.

Tables XXXIII through XXXVI represent current estimates of factory costs for producing the tunnel diode amplifiers in the current design. Two significant cost changes have been made. First the tunnel diode costs have increased 50%, due to yield forecasts and increased conservatism on the part of the Semiconductor Products Department. Second, the cost estimate on the enclosure represents current quotes on casting and finishing. Higher volume costs are not shown for 1971 and 1973 because production facilities are inadequate to provide these volumes.



(a)
Interior Details



(b)
Exterior and Mount

Figure 64. TDA Package

TABLE XXXIII
TUNNEL DIODE AMPLIFIER - FACTORY COST ESTIMATE

1971

TDA		10^3/yr.
MATERIALS		
TD Substrate		60.00
Circulator SS		.62
4 Si Resistors		5.00
2 Ferrite Discs		1.05
4 Bar Magnets		.46
Enclosure		9.00
2 Connectors		5.66
Potentiometer		.80
Zener Diode		.16
Feedthru Capacitor		.16
Cover Plate		.30
Misc.		.10
		83.31
Freight and Spoilage (5%)		4.16
		87.47
Labor	Wire Shop	10.50
	Photolithography	2.20
	Test	.25
TOTAL		$100.42

TABLE XXXIV

TUNNEL DIODE AMPLIFIER - FACTORY COST ESTIMATE

1973

TDA		10^3	10^4	10^5
MATERIALS				
TD Substrate		45.00	30.00	15.00
Circulator SS		.66	.12	.06
4 Si Resistors		4.60	3.20	2.60
2 Ferrite Discs		1.11	1.06	1.00
4 Bar Magnets		.49	.34	.25
Enclosure		9.55	8.50	7.50
2 Connectors		6.00	4.25	3.03
Potentiometer		.85	.80	.80
Zener Diode		.16	.16	.16
Feedthru Capacitor		.17	.17	.17
Cover Plate		.32	.25	.20
Misc.		.12	.12	.12
		69.03	48.97	30.89
Freight & Spoilage		3.45	2.45	1.55
Labor	Wire Shop	11.13	1.50	60
	Photolith	2.32	1.50	1.00
	Test	27	20	15
TOTAL		$86.20	$54.62	$34.19

TABLE XXXV

TUNNEL DIODE AMPLIFIER - FACTORY COST ESTIMATE

1975

TDA		10^3	10^4	10^5	10^6
MATERIALS					
TD Substrate		22.50	16.50	12.00	10.50
Circulator SS		.70	.13	.06	.05
4 Si Resistors		4.00	3.00	2.40	2.00
2 Ferrite Discs		1.18	1.12	1.07	1.00
4 Bar Magnets		.52	.36	.27	.25
Enclosure		10.10	9.00	7.90	6.75
2 Connectors		6.37	4.50	3.23	2.74
Potentiometer		.90	.84	.84	.84
Zener Diode		.16	.16	.16	.16
Feedthru Cap.		.18	.18	.18	.18
Cover Plate		.34	.27	.21	.20
Misc.		.13	.13	.13	.13
		47.08	36.19	28.45	24.80
Freight & Spoilage		2.36	1.81	1.43	1.24
Labor	Wire Shop	11.80	1.60	.64	.50
	Photolith	2.48	1.60	1.12	1.00
	Test	.28	.21	.16	.10
TOTAL		$64.00	$41.41	$31.80	$27.64

TABLE XXXVI

TUNNEL DIODE AMPLIFIER - COST SUMMARY

TDA	10^3	10^4	10^5	10^6
1971	100.42	-	-	-
1973	86.20	54.62	34.19	-
1975	64.00	41.41	31.80	27.64

6.10 ADDENDUM

Additional Effort on the T.D.A. Redesign

Purpose -

The purpose of this section is to describe the follow-up effort (G. E. internal funding) to improve the stability and matching, and, hence, gain and bandwidth of the TDA.

Analysis -

Typical device characteristics of the X-Band microwave tunnel diodes, which can be expected to have reasonable yields, are:

R_s = 5.7 ohms

L_s = 0.25 nanohenries

C_d = 0.20 picofarads

R_d = -71.0 ohms

where R_s is the series resistance, L_s is the series inductance, C_d is the diode junction capacitance, and R_d is the negative resistance. Then the calculated admittances of the assumed tunnel diodes are:

Freq. in GHz	Admittance in Mhos		
11.	-0.0242	+j	0.0145
11.2	-0.0246		0.0148
11.4	-0.0251		0.0151
11.6	-0.0256		0.0155
11.8	-0.0262		0.0158
11.9	-0.0265		0.0160

Freq. in GHz (cont'd)	Admittance in Mhos (cont'd
12.0	-0.0267 +j 0.0162
12.1	-0.0270 0.0163
12.2	-0.0273 0.0165

From this it can be seen that, in order to insure stability, it is necessary to have the load conductance across this diode larger than 0.03 mhos, preferably with positive susceptance to prevent potential oscillation. At the same time, in order to have good gain, by reflection, the design should present a proper negative conductance to the circulator using matching sections.

To design such a matching section involves handling negative conductances; and an extension of the Smith Chart to cover both positive and negative conductances is very helpful. Since no such chart exists, the following adaptation was made.

As usual, the reflection coefficient is defined by

$$\Gamma = \frac{Y_o - Y_L}{Y_o + Y_L} = x + j\,y \quad .$$

Then the normalized quantities are:

$$G' = G_L/Y_o = \frac{1 - x^2 - y^2}{1+2x+x^2+y^2}$$

$$B' = B_L/Y_o = \frac{-2y}{(1+x)^2 + y}$$

From these the constant G' and B' circles are,

$$(x + G'/(G' + 1)\,)^2 + y^2 = 1/(G' + 1)^2$$

$$(x + 1)^2 + (y + 1/B')^2 = (1/B')^2$$

Set y = 0, for constant G' circles; the x-axis intercepts are

$$x = -1 \ \& \ \frac{1 - G'}{1 + G'} \quad ; \qquad \text{Radius} = \left| \frac{1}{G' + 1} \right|$$

Set x = -1; for constant B' circles, the y-axis intercepts at 0 and -2/B' and the radius is equal to $|1/B'|$. With these relations, we calculate the various intercepts and radius and plot the extended Smith Chart shown in Figure 65. It is interesting to note that the G' = -1 value is represented by a vertical line, x = -1. The circles, which have G' between 0 and -1, are beyond the normal Smith Chart unit circle but on the right side of the x = -1 line. All the circles, with G' -1 are on the left side of this line. The constant B' circles are just the continuation of the normal Smith Chart.

With this extended Smith Chart we can find the matching by conventional procedures, such as rotating the Γ vector along a transmission line, adding susceptances, etc. We soon find that the configuration shown in Figure 66 satisfies our requirements. The point, P_1, is the diode admittance. An inductive susceptance added to P_1 brings it to point, P_2. A short section of transmission line after that brings it to P_3. Adding a proper positive susceptance brings it to P_4. The reflection coefficient corresponding to the admittance at P_4 gives the gain (=20 log $|\Gamma|$, in dB.).

Looking from the TD terminals, with the help of the chart, we can also verify that the load satisfies the requirement for stability, i.e. the positive real part of the admittance is larger than 0.03 mhos, or 1.5, when normalized to the 50-ohm line.

The exact line lengths of the matching sections are determined by computation over the frequency band. A final design printout is shown in Table XXXVII, where the last column gives the gain before allowing for any other losses. D_1 through D_4, in inches, are the line lengths used for the design using 1/32" teflon glass boards in a strip-line configuration. Although the components are ready to be assembled no experimental results can be reported.

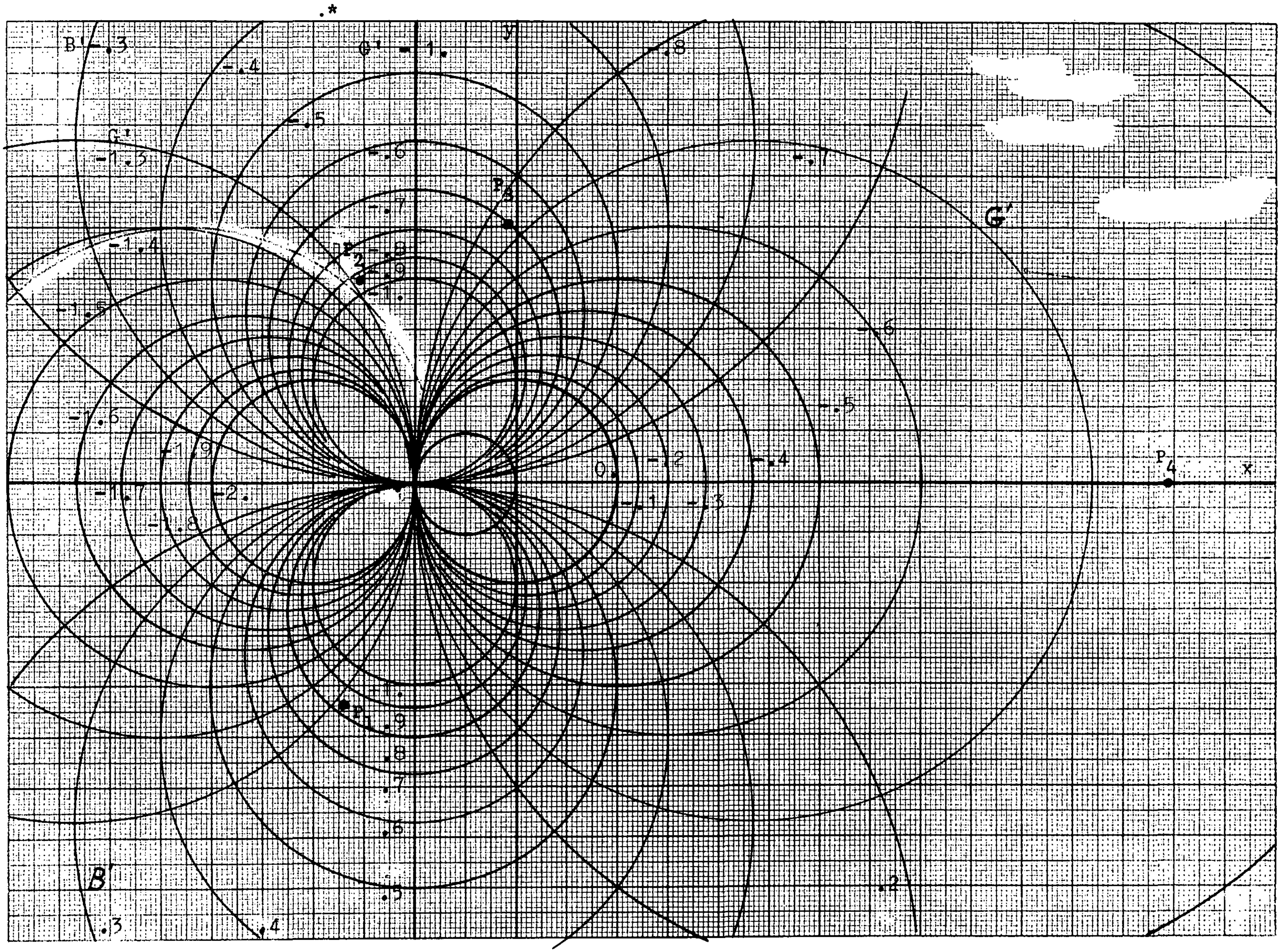
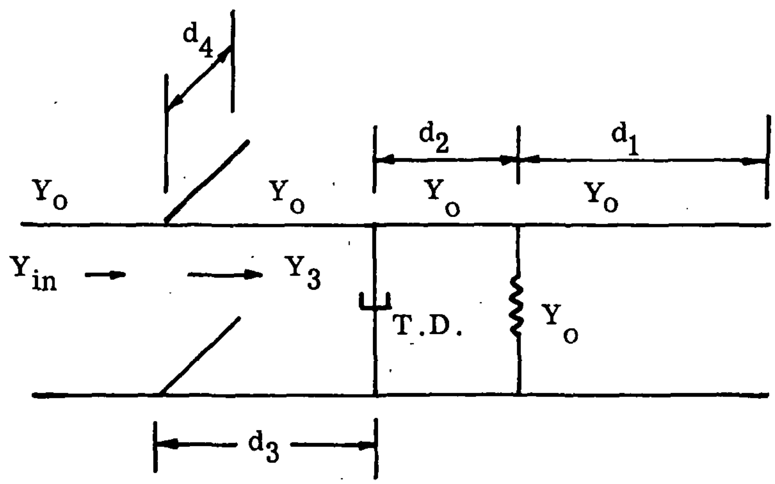


Figure 65. Designing of TDA Matching on Extended Smith Chart Covering Negative Conduction



$Y_o = 0.020$ mhos

Figure 66. T.D.A. Matching and Stabilizing Configuration

TABLE XXXVII

TUNNEL DIODE AMPLIFIER - FINAL DESIGN PRINTOUT

TDA2 13:37EDT 05/23/72

D3=?.0458

D4=?.048

D1=?.1545

D2=?.065

FREQ(GHZ)	Y IN (MHØ)		Y3		GAMMA	PØWER GAIN	DB.
10.500	-0.007	-0.007	-0.007	-0.016	1.336	0.939	4.260
10.550	-0.007	-0.007	-0.007	-0.016	1.408	0.934	4.555
10.600	-0.007	-0.007	-0.007	-0.016	1.482	0.926	4.851
10.650	-0.008	-0.006	-0.008	-0.015	1.559	0.917	5.149
10.700	-0.008	-0.006	-0.008	-0.015	1.639	0.905	5.448
10.750	-0.008	-0.005	-0.008	-0.015	1.721	0.891	5.748
10.800	-0.009	-0.005	-0.009	-0.014	1.807	0.874	6.051
10.850	-0.009	-0.005	-0.009	-0.014	1.895	0.854	6.354
10.900	-0.009	-0.004	-0.009	-0.014	1.986	0.831	6.660
10.950	-0.010	-0.004	-0.010	-0.014	2.080	0.805	6.967
11.000	-0.010	-0.004	-0.010	-0.013	2.177	0.776	7.276
11.050	-0.010	-0.003	-0.010	-0.013	2.277	0.742	7.587
11.100	-0.011	-0.003	-0.011	-0.013	2.381	0.704	7.900
11.150	-0.011	-0.003	-0.011	-0.012	2.488	0.662	8.214
11.200	-0.011	-0.002	-0.011	-0.012	2.599	0.614	8.531
11.250	-0.012	-0.002	-0.012	-0.012	2.712	0.561	8.849
11.300	-0.012	-0.002	-0.012	-0.012	2.830	0.503	9.170
11.350	-0.012	-0.001	-0.012	-0.011	2.951	0.438	9.493
11.400	-0.013	-0.001	-0.013	-0.011	3.075	0.366	9.818
11.450	-0.013	-0.001	-0.013	-0.011	3.203	0.287	10.146
11.500	-0.013	-0.001	-0.013	-0.011	3.334	0.200	10.475
11.550	-0.014	-0.000	-0.014	-0.010	3.469	0.105	10.807
11.600	-0.014	0.000	-0.014	-0.010	3.607	-0.000	11.142
11.650	-0.014	0.000	-0.014	-0.010	3.747	-0.115	11.479
11.700	-0.015	0.001	-0.015	-0.010	3.891	-0.241	11.818
11.750	-0.015	0.001	-0.015	-0.010	4.037	-0.379	12.160
11.800	-0.015	0.001	-0.015	-0.009	4.186	-0.530	12.504
11.850	-0.016	0.001	-0.016	-0.009	4.336	-0.695	12.851
11.900	-0.016	0.001	-0.016	-0.009	4.487	-0.874	13.201
11.950	-0.016	0.002	-0.016	-0.009	4.638	-1.070	13.552
12.000	-0.017	0.002	-0.017	-0.009	4.789	-1.284	13.906
12.050	-0.017	0.002	-0.017	-0.009	4.938	-1.516	14.263
12.100	-0.017	0.002	-0.017	-0.009	5.085	-1.768	14.621
12.150	-0.018	0.002	-0.018	-0.008	5.227	-2.041	14.981
12.200	-0.018	0.003	-0.018	-0.008	5.363	-2.337	15.344
12.250	-0.018	0.003	-0.018	-0.008	5.491	-2.658	15.707
12.300	-0.019	0.003	-0.019	-0.008	5.608	-3.004	16.072
12.350	-0.019	0.003	-0.019	-0.008	5.712	-3.377	16.438
12.400	-0.019	0.003	-0.019	-0.008	5.799	-3.778	16.804

PRØGRAM STØP AT 750

USED .96 UNITS

7.0 EVALUATION OF PROTOTYPE CONVERTERS

Each of the 30 prototype converters fabricated during the program was subjected to quantitative and qualitative tests on performance.

7.1 DESIGN REQUIREMENTS

All of the design requirements for the converters were met with the prototype units.

The converter types fabricated were designed to operate with the following input signal formats.

Type A. 2.25 GHz, vestigial side-band modulation

Type B. 2.25 GHz, frequency modulation with a modulation index of two.

Type D. 12.0 GHz, frequency modulation with a modulation index of three.

Each converter type was constructed in two units, one antenna mounted and the other located at the television receiver. The interconnecting cable used in each case was 300 ohm twinlead and the converters were tested using a 30 foot cable length between units.

The converters were operated with a conventional NTSC color television receiver in an unmodified condition. The RF interface between the converter and the television receiver was completed using a twinlead connection with the signal generated by the converters occupying the standard channel six frequency allocation.

The source of power for the converters was 115 VAC, which was converted into convenient dc voltages to supply the indoor unit circuits and the antenna unit circuit via the twinlead interconnection.

All converters used power transformers with primary fusing for protection and isolation. The dc voltages on the twinlead were less than 24 VDC in all cases to prevent electrical hazard to the user.

A standard NTSC color signal format was used as a modulating composite video source. The sound subcarrier was added to the composite video signal prior to modulation at a reduced level from that normally used for terrestial AM-VSB transmission. The output of the converter was designed to be a AM-VSB signal at RF in the Type A converter and an AM-DSB signal for the FM converters. This latter format did not affect the television receiver operation because of the receiver attenuation of the unused sideband. The AM-DSB signal format could not interfere with the TV receiver operation from normal stations because the converters were switched off for this mode of operation.

The converters were operated over the following ranges of antenna output signal level.

Type A. -110 dBW to -80 dBW at sync peak
Type B. -112 dBW to -92 dBW
Type D. -110 dBW to -90 dBW

An antenna output impedance of 50 Ω was assumed in the design. The antenna unit output impedance and receiver unit input and output impedances were maintained at 300 Ω, balanced to interface with the twinlead and television receiver terminals.

The weight, size and power consumption of the converters are listed in Table XXXVIII. These values are compatible with the installation and operational objectives.

TABLE XXXVIII

CONVERTER AND WEIGHT POWER VALUES

	Antenna Unit		Indoor Unit		Total		Input
	Kg.	Oz.	Kg.	Oz.	Kg.	Oz.	Power (Watts)
X-Band FM	0.62	22	1.30	46	1.87	66	9
S-Band FM	0.40	14	1.13	40	1.53	54	5
S-Band AM	0.91	32	0.76	27	1.67	59	5

7.2 DESIGN OBJECTIVES

The design objectives of the contract concerned the following areas.

1) Degradation of television receiver performance caused by the converter interconnection
2) Intermodulation
3) Bandwidth and Envelope Delay
4) Signal to Noise Ratio at the TV receiver display.
5) Environmental specifications.

All design objectives were met with the exception of signal-to-noise ratio for specific units. A major portion of the converter testing was conducted to satisfy performance objectives over the specified temperature range of -40°C to +55°C for antenna mounted units and -1°C to 60°C for receiver mounted units. The most difficult performance area for all converters was the generation of local oscillator power with temperature stable power level and frequency.

With regard to the specific performance objectives the following results were obtained.

7.2.1 Degradation of Television Receiver Performance

The converters were designed in each case with antenna signal switches to permit selection of the television receiver signal source either from the converter being used or the normal terrestial antenna system. When the local antenna system is used to feed the television receiver the converter is turned off by switching its prime ac power off. The only performance degradation to the receiver is a slight signal attenuation resulting from routing the local antenna signal through the switch. This attenuation was computed to be less than one decibel at the highest channel frequency in the VHF band, 220 MHz.

7.2.2 Intermodulation

The intermodulation specification placed on the converters was that the signal to intermodulation distortion ratio be greater than 40 dB in both the picture and sound channels. This specification is difficult to verify quantitatively for two reasons. First, the contribution due to the television receiver itself cannot be isolated within the video and sound channels of the receiver and, second, it is dependent on television receiver fine tuning and picture controls. A substitution method of qualitative evaluation was used to establish compliance with this design objective. The procedure used was as follows.

A known level of 920 kHz signal, the most predominant intermodulation product in color television processing, was injected at a level that was 40 dB below the video level in the FM converter demodulation. The resulting display was then compared with the picture quality obtained with the converter alone. With proper IF and discriminator frequency alignment there was no discernable 920 kHz component visible on the television picture tube. The added signal at 40 dB below the video resulted in a visible but not too objectionable intermodulation type herringbone pattern.

The sound channel was evaluated qualitatively in that no noticeable 60 Hz hum was produced with proper converter alignment. This intermodulation could be detected when excessive sound subcarrier signal level was used in the FM signal modulation or when the converter was misaligned.

The AM converters did not produce any noticeable intermodulation products.

7.2.3 Bandwidth and Envelope Delay

The bandwidth of the video processing circuits in the FM converters was purposely designed to have excess bandwidth to eliminate any differential phase delay in the video signals. The design philosophy was to let the television receiver determine the signal bandwidth and delay characteristics of the system. The FM converter IF amplifier and discriminator circuits were operated with sufficient bandwidth to assure the fidelity of the signals. The discriminator linearity could introduce some differential gain errors. A balanced discriminator circuit was selected to minimize this error source. The AM remodulator circuit was operated at a low power level to insure linearity and minimum spurious radiation.

The AM converter system was designed as a relatively wideband frequency translator. There was negigible gain variation across the channel bandwidth by the one tuned amplifier used in this converter design.

7.2.4 Signal-to-noise Ratio

The one design objective that was not met in all cases was the output signal-to-noise ratio.

The design goals for the output signal to noise were as follows:

Type A. 35 dB S/N for -92 dBW antenna output power level.

Type B. 35 dB S/N for -107 dBW antenna output power level.

Type D. 35 dB S/N for -105 dBW antenna output power level.

The greatest variation of output signal to noise ratio was encountered in the AM converter systems. The cause of this variation was found to be a mismatch at the mixer RF input port. The performance of those converters that exhibited lower output S/N was greatly improved by using a stub tuner in the RF input line. A further diagnosis of the problem was prevented by integrated stripline assembly of the mixer, local oscillator filter and step recovery diode circuits. It is believed that this problem is caused by the L. O. filter and mixer interface, because it does not exist in the S-Band FM converters which use an identical mixer without the L. O. filter.

As previously mentioned, the output S/N of the Type B converters, S-Band FM, is more consistent. The output signal-to-noise ratio is established predominantly by the converter noise figure, which, in turn, depends on the mixer diode characteristics and the noise performance of the first IF amplifiers. The converter performance variations from unit to unit indicate that a higher quality mixer diode and/or IF amplifier transistor is required to meet the output S/N objective in production. The best of the 10 Type B converters meet the output S/N objective but the yield is inadequate.

The performance of the Type D, X-Band FM converters is the best in terms of output S/N at nominal RF input signal level. This can be attributed to two factors, first a high quality mixer diode was used as a result of a limitation in available alternatives and, second, the wider modulation index used in this system provides an excess of FM signal-to-noise improvement over that required to meet the output S/N objective.

The improved consistency of the X-Band FM converter noise figures from unit to unit over the S-Band FM converters, tends to indicate that the S-Band FM mixer diodes are the prime cause of noise figure variance in the Type B converters. This deduction is based on the fact that identical IF amplifiers are used in both types of FM converters. It is still reasonable that a better IF transistor should be used in the first IF stages. Both FM systems could have their average noise figures (unit to unit) reduced by at least 2 dB through the substitution of a lower noise first IF transistor than the ten cent unit selected for the prototype converters.

7.2.5 Temperature Tests

As previously mentioned, the most difficult performance parameter concerned the operation of the antenna mounted units over the range of -40°C to +55°C. The most difficult function to achieve was the generation of microwave local oscillator power that was frequency stable and power level stable in minimum cost configurations.

The frequency stability of the Type A converter local oscillator was established by using a crystal reference. The temperature problem with this approach was concerned with obtaining a multiplier chain between the crystal reference oscillator and the desired microwave power. The multiplication was performed in two steps, first with a class C transistor tripler and second, with a times — ten step recovery diode multiplier. The conversion efficiency of the transistor tripler is dependent on the class C stage conduction angle and the tuning at the tripler output. Variation of these parameters with temperature could result in both power level changes at the desired output frequency and generation of unwanted harmonics of the crystal reference frequency. The step-recovery diode circuit was less sensitive to temperature when driven from a fixed resistive source. A stable combination was achieved by buffering the triple output and inserting a resistive loss pad between the tripler buffer and the step-recovery diode input. Table XXXIX shows the results of temperature tests made on the Type A prototype unit local oscillators.

In all cases the local oscillator frequency stability was as desired and the output power level of the local oscillator was within desired limits. The tolerance to line voltage variation reflects the alignment stability and the performance of the power supply voltage regulator.

The FM converters presented a different problem with regard to temperature variations. The reference oscillator - multiplier approach was tried originally, as in the Type A converter, to generate frequency stable local oscillator power. The suppression of undesired harmonic content in the FM converters was more critical and adequate performance could not be obtained over the temperature range with reasonable circuit complexity. The approach in both FM converters was to generate the local oscillator power with oscillators operating directly at the desired frequency. This approach was possible because the frequency stability requirements of the local oscillators in the FM systems could be relaxed from that required in the direct translation Type A converter. Temperature compensated oscillators were designed for both FM converter types. Tables XL and XXXVI provide representative data on the Type B converter transistor local oscillator frequency stability. Table XL shows the uncompensated oscillator. The oscillator bias voltage versus temperature characteristic is obtained to give a measure of the amount of compensation required by the thermistor in the oscillator bias circuit, Table XLI shows the results of the compensation, i. e., compensated frequency versus temperature. The worst case frequency drift occurs in converter #7 and 80% of this variation occurs below -17.8°C. A remote local oscillator frequency adjustment is provided in the Type B converter indoor units. A potentiometer is provided to adjust the bias voltage fed to the antenna unit and this adjustment can be used to set the local oscillator frequency.

TABLE XXXIX.

TYPE A CONVERTER, TEMPERATURE TEST OF LOCAL OSCILLATOR

Unit #		Temperature -40°C	20°C	+55°C	Minimum Tolerable Line voltage to Converter -40°C	+55°C
1	Frequency, f_{LO} (GHz)	2.1671	2.1671	2.1671		
	L.O. Power at Ant. Port.[1] (dBm)	-25.7	-26.0	-27.2	105 vac	92 vac
	Spurious below f_{LO} (dB)	-28	-26	-24		
2		2.1672	2.1673	2.1673		
		-25.8	-26.3	-27.3	82	94
		-36	-38	-35		
3		2.1672	2.1672	2.1673		
		-24.3	-26.3	-27.5	98	92
		-36	-26	-26		
4		2.1673	2.1674	2.1674		
		-25.9	-26.2	-27.0	84	93
		-35	-30	-28		
5		2.1672	2.1673	2.1673		
		-23.8	-26.3	-27.2	90	95
		-31	-34	-32		
6		2.1672	2.1672	2.1672		
		-25.0	-26.4	-27.7	88	90
		-31	-28	-26		
7		2.1674	2.1674	2.1674		
		-25.4	-26.8	-27.8	85	100
		-30	-30	-29		
8		2.1673	2.1674	2.1673		
		-26.7	-27.3	-28.4	96	104
		-25	-28	-28		
9		2.1671	2.1672	2.1672		
		-24.4	-26.3	-28.2	92	96
		-26	-29	-28		
10		2.1675	2.1675	2.1675		
		-26.4	-25.9	-27.0	92	94
		-29	-27	-29		

(1) Measurement includes 20 dB attenuator loss used in test.

TABLE XL.

TYPE B CONVERTER - L.O. TEMPERATURE TESTS

Bias Voltage Required to Hold Frequency Constant

Serial No.	$-40^{o}C$	0	+20	+55
2	10.1	11.1	11.8	13.8
5	13.8	13.95	13.5	14.2
7	9.8	10.8	11.2	12.1
9	9.2	12.0	13.2	14.9

TABLE XLI.

FREQUENCY VERSUS TEMPERATURE AT CONSTANT VOLTAGE AFTER COMPENSATION

	Frequency in MHz				
Serial No.	$40^{o}C$	0	+20	+55	$\lvert\Delta F\rvert$ (MHz)
2	2124.46	2126.16	2130.07	2125.60	6.41
5	2129.12	2128.81	2130.04	2128.00	2.04
7	2138.13	2129.35	2130.08	2128.13	10.00
9	2133.65	2130.67	2130.02	2129.55	4.1

The Type D converter local oscillator is a temperature compensated Gunn diode oscillator. The frequency versus temperature of the oscillators in the prototype converters is listed in Table XLII. The majority of these units have excellent temperature and power level stability over the temperature range.

Units #2, 4 and 9 have marginal temperature stability. The characteristics of these units are still adequate to provide good performance. Additional product design is required in the Gunn oscillator mechanical construction to improve temperature stability and tuning ease. Several variations were investigated for mounting the tuning probe during the development but none proved to be entirely satisfactory.

Temperature Tests were also conducted on the indoor units of each converter type. No problems were encountered because of the less stringent temperature range, $-1.1^{O}C$ to $+60^{O}C$, and because the circuits were less susceptible to ambient temperature variations.

7.2.6 Spurious Radiation

Two possible sources of spurious radiation exist in each converter type. The most serious problem is that of local oscillator radiation from the ground receiver unit antennas. In the Type A converters the L.O. level at the antenna port of the mixer front end averages about -6 dBm. This level is 86 dB above the nominal received signal level and could present an operational problem in high density installations. The problem is relieved somewhat by antenna directivity, and relative antenna-to-antenna orientation required for proper operation minimizes the interantenna coupling. The best solution for minimizing this potential problem is the use of RF amplffiers in the Type A and Type B ground receiving systems. This approach was not used in the present converters because of the relative cost increase of including them.

The radiated local oscillator power from the Type D converters is higher than from the Type A converter. The average local oscillator power level at the antenna port of the mixer is about 0 dBM. This increase in power is offset by increased path loss at the 12.0 GHz frequency between antennas. The FM systems are more susceptible to IF overload because of the higher IF gain, and adjacent converters could produce overload signals in one or both IF amplifiers at a frequency equal to their local oscillator frequency difference. Elimination of this potential problem is less readily achieved with a RF preamplifier and is even more difficult with a RF filter because of the higher frequency. A preamplifier or circulator between the antenna and mixer would provide the best solutions.

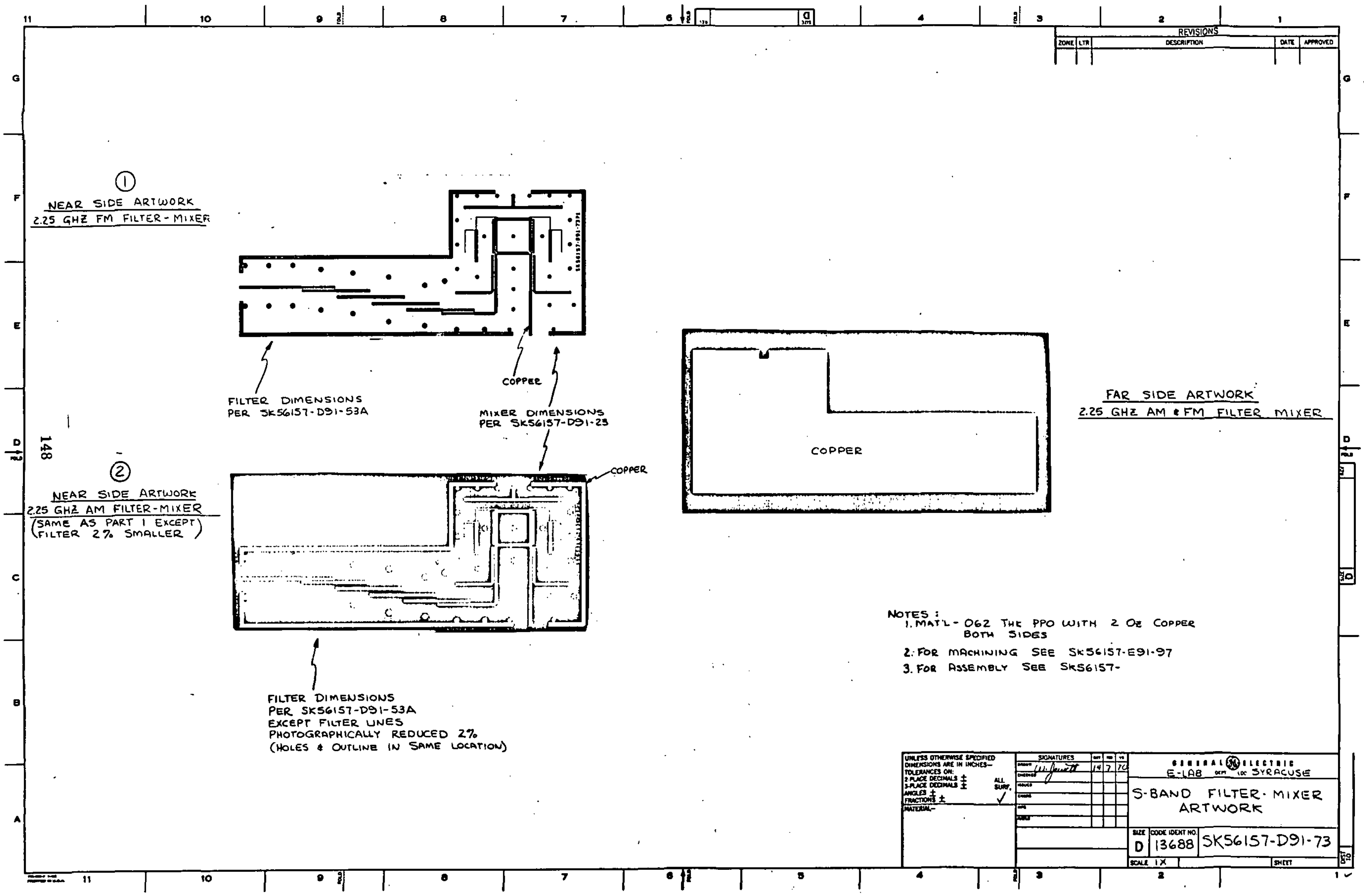
The other potential source of spurious radiation is the remodulator circuits in the FM converters. This has been minimized by operating at low levels in the remodulator circuits and by removing converter power when local stations are received. There was no indication of mutual interference from converter to adjacent television receivers in the same room when testing the FM converters.

TABLE XLII.

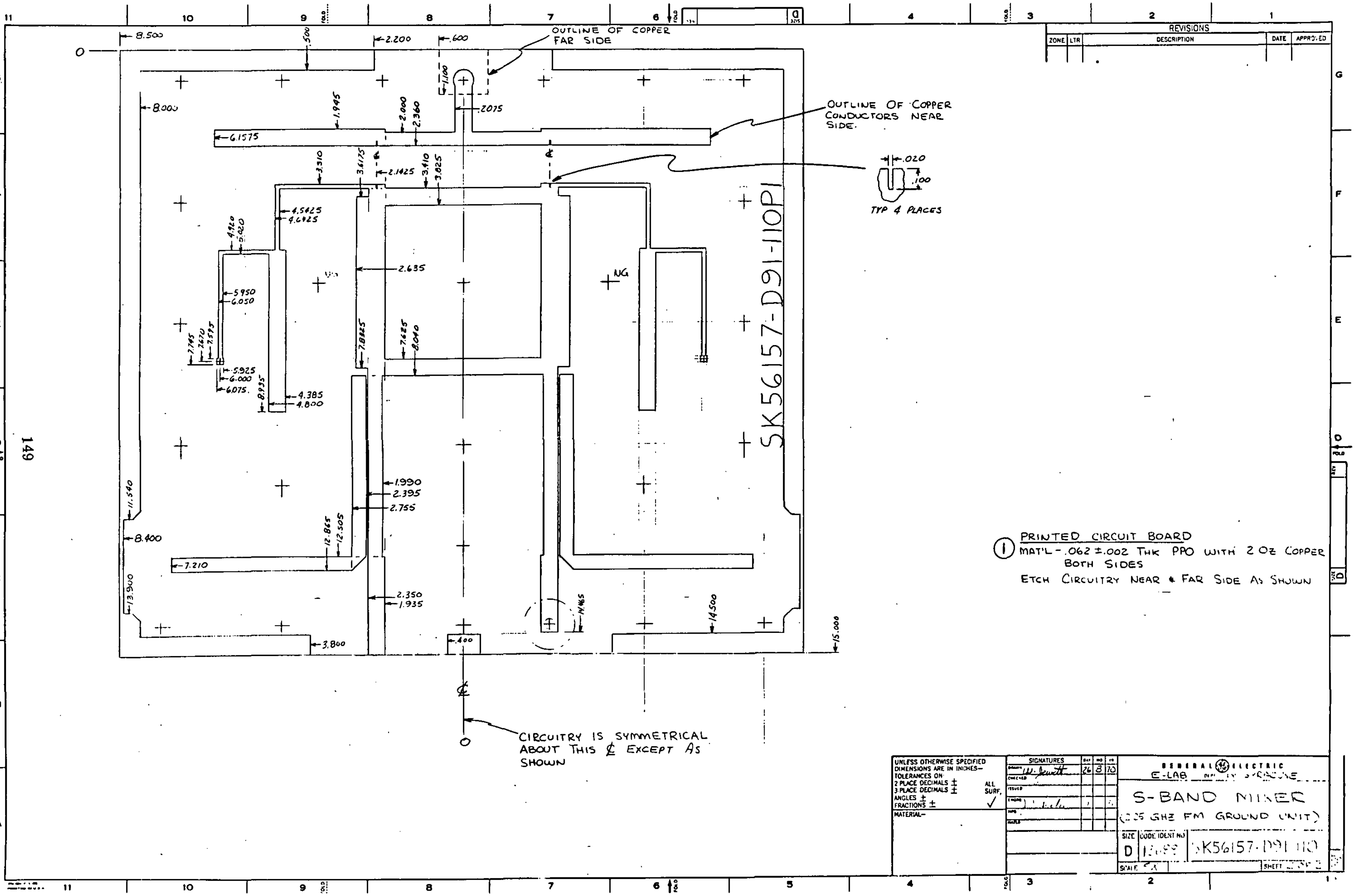
TYPE D CONVERTER TEMPERATURE TEST OF LOCAL OSCILLATOR

Serial No.	L. O. Frequency and Power*				|ΔF|
	Ambient Temperature				MHz
	-40°C	-10°C	20°C	55°C	
2	11.8789 -20.4	11.8760 -20.3	11.8765 -21.0	11.8713 -21.4	7.6
3	11.8700 -18.3	11.8703 -18.4	11.8703 -18.9	11.8711 -19.3	1.1
4	11.8705 -21.05	11.8723 -21.9	11.8806 -22.4	11.8816 -22.6	11.1
5	11.7650 -19.5	11.7670 -20.5	11.7691 -20.6	11.7693 -21.4	4.3
6	11.8798 -21.7	11.8816 -22.3	11.8829 -22.7	11.8840 -23.9	4.2
7	11.8970 -20.8	11.8993 -21.5	11.8998 -22.3	11.9000 -22.8	3.0
8	11.9056 -18.1	11.9058 -18.6	11.9071 -18.8	11.9084 -19.7	2.8
9	11.8839 -17.4	11.8834 -17.8	11.8818 -18.0	11.8778 -17.7	6.1
10	11.8906 -22.1	11.8932 -22.4	11.8923 -22.7	11.8923 -23.3	2.6

*L. O. Power measurement includes 20 dB attenuator used in test.

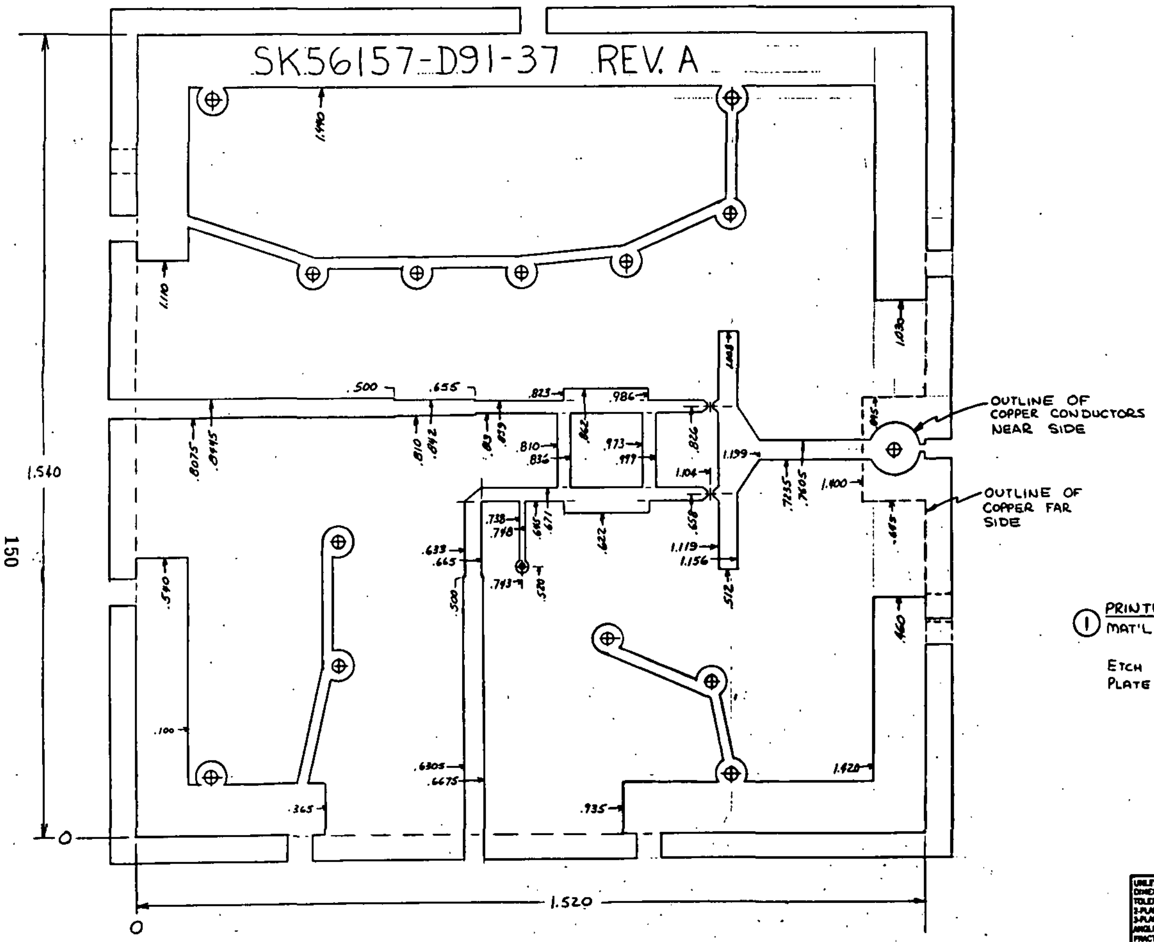


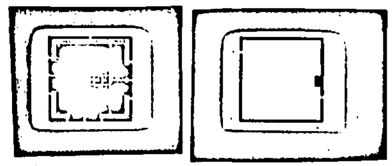
1
NEAR SIDE ARTWORK
2.25 GHZ FM FILTER-MIXER
FILTER DIMENSIONS
PER SK56157-D91-53A
COPPER
MIXER DIMENSIONS
PER SK56157-D91-23
2
NEAR SIDE ARTWORK
2.25 GHZ AM FILTER-MIXER
(SAME AS PART 1 EXCEPT
FILTER 2% SMALLER)
COPPER
FILTER DIMENSIONS
PER SK56157-D91-53A
EXCEPT FILTER LINES
PHOTOGRAPHICALLY REDUCED 2%
(HOLES & OUTLINE IN SAME LOCATION)
COPPER
FAR SIDE ARTWORK
2.25 GHZ AM & FM FILTER MIXER
NOTES:
1. MAT'L - .062 THK PPO WITH 2 OZ COPPER BOTH SIDES
2. FOR MACHINING SEE SK56157-E91-97
3. FOR ASSEMBLY SEE SK56157-
REVISIONS
ZONE
LTR
DESCRIPTION
DATE
APPROVED
UNLESS OTHERWISE SPECIFIED
DIMENSIONS ARE IN INCHES—
TOLERANCES ON:
2 PLACE DECIMALS ±
3 PLACE DECIMALS ±
ANGLES ±
FRACTIONS ±
MATERIAL—
ALL SURF.
SIGNATURES
14 7 70
GENERAL ELECTRIC
E-LAB
SYRACUSE
S-BAND FILTER-MIXER
ARTWORK
SIZE
CODE IDENT NO.
D
13688
SK56157-D91-73
SCALE 1X
SHEET



OUTLINE OF COPPER FAR SIDE
OUTLINE OF COPPER CONDUCTORS NEAR SIDE.
.020
.100
TYP 4 PLACES
5K56157-D91-110P1
NG
CIRCUITRY IS SYMMETRICAL ABOUT THIS ℄ EXCEPT AS SHOWN
① PRINTED CIRCUIT BOARD
MAT'L - .062 ±.002 THK PPO WITH 2 OZ COPPER BOTH SIDES
ETCH CIRCUITRY NEAR & FAR SIDE AS SHOWN
REVISIONS
ZONE LTR DESCRIPTION DATE APPROVED
UNLESS OTHERWISE SPECIFIED DIMENSIONS ARE IN INCHES—
TOLERANCES ON
2 PLACE DECIMALS ±
3 PLACE DECIMALS ±
ANGLES ±
FRACTIONS ±
ALL SURF. ✓
MATERIAL—
SIGNATURES
26 8 70
GENERAL ELECTRIC
E-LAB
S-BAND MIXER
GROUND UNIT
SIZE D
CODE IDENT NO
5K56157-D91
SCALE
SHEET

ZONE	LTR	REVISIONS DESCRIPTION	DATE	APPROVED
	A	REDRAWN ADDING MTG. HOLES, PAD LINKS, OUTPUT PAD TO PREAMP, & GOLD PLATING. INCREASED HEIGHT & WIDTH. CIRCUITRY UNCHANGED	12 MAY 1970	W. Jewett



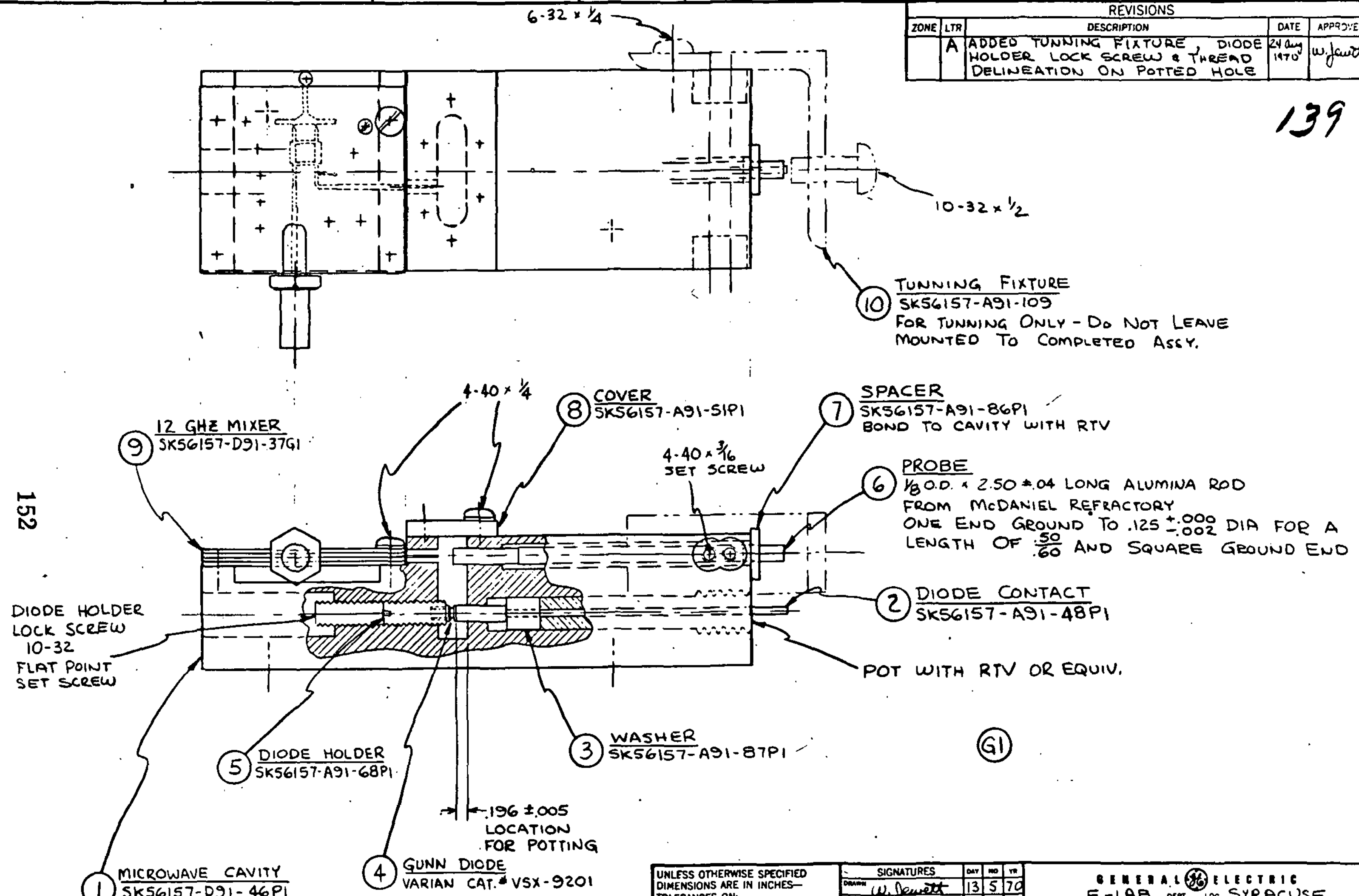


① PRINTED CIRCUIT BOARD
MAT'L - 1/32 THK PPO WITH 2 OZ COPPER BOTH SIDES.
ETCH CIRCUITRY NEAR & FAR SIDES AS SHOWN.
PLATE BOTH SIDES WITH .0001 THICK SELREX CI GOLD.

UNLESS OTHERWISE SPECIFIED DIMENSIONS ARE IN INCHES— TOLERANCES ON: 2-PLACE DECIMALS ± 3-PLACE DECIMALS ± ANGLES ± FRACTIONS ± MATERIAL— ALL SURF. ✓	SIGNATURES: W. Jewett 3/10/69; [signature] 6/7/70	GENERAL ELECTRIC E-LAB SYRACUSE
		12 GHZ MIXER
		SIZE D — CODE IDENT NO. 13688 — SK56157-D91-37
		SCALE 10X — SHEET 2 OF 2

REVISIONS				
ZONE	LTR	DESCRIPTION	DATE	APPROVED
	A	ADDED TUNNING FIXTURE, DIODE HOLDER LOCK SCREW & THREAD DELINEATION ON POTTED HOLE	24 Aug 1970	W. Jewett

139



UNLESS OTHERWISE SPECIFIED DIMENSIONS ARE IN INCHES—
TOLERANCES ON:
2-PLACE DECIMALS ±
3-PLACE DECIMALS ±
ANGLES ±
FRACTIONS ±
ALL SURF. ✓
MATERIAL—

SIGNATURES		DAY	MO	YR
DRAWN	W. Jewett	13	5	70
CHECKED				
ISSUED				
ENGRG	J. Heller	24	8	70
MFG				
RELEASE				

GENERAL ELECTRIC
E-LAB DEPT LOC SYRACUSE

MIXER - CAVITY ASSY.
(12GHZ FM ANTENNA UNIT)

SIZE	CODE IDENT NO.	
C	13688	SK56157-C91-96

SCALE 2X SHEET

FN 903-F PRINTED IN U.S.A.

The Electronics Laboratory is the center of General Electric's applied research and advanced development activities in electronics and related fields. It is a component of the Company's Electronic Systems Division. The work of the Laboratory is oriented primarily in support of General Electric's present and anticipated product interests, particularly in the defense and industrial markets. Engineers and scientists from a variety of disciplines work together in programs that continually require state-of-the-art advances. By maintaining a close working relationship with other scientific and technological centers within the Company, the staff is able to apply the full resources of the General Electric Company to any Electronics Laboratory program.

Progress Is Our Most Important Product

GENERAL ELECTRIC